# AN IMPROVED COMMUNICATIONS SYSTEM FOR FREE SWIMMING DIVERS

Peter Stewart Pierpont

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REPORT

# NAVAL POSTGRADUATE SCHOOL

## Monterey, California

AN IMPROVED COMMUNICATIONS SYSTEM
FOR FREE SWIMMING DIVERS

by

Peter Stewart Pierpont

December 1976

Thesis Advisor: George H. Marmont

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In a continuing effort to produce an efficient, reliable communications system for free swimming divers, a second-generation, frequency modulated underwater communications system was designed, analyzed and tested. This system is given the acronym DUCS-II (DIVER UNDERWATER COMMUNICATIONS SYSTEM), and incorporates many features of a prototype (DUCS I) previously designed at the Naval Postgraduate School. Improvements or alternative methods concerning electronic circuitry, underwater enclosures

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An Improved Communications System
For Free Swimming Divers

by

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requirements of the degree of

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## ABSTRACT

In a continuing effort to produce an efficient, reliable communications system for free swimming divers, a second-generation, frequency modulated underwater communications system was designed, analyzed and tested. This system is given the acronym DUCS-II (DIVER UNDERWATER COMMUNICATIONS SYSTEM), and incorporates many features of a prototype (DUCS I) previously designed at the Naval Postgraduate School. Improvements or alternative methods concerning electronic circuitry, underwater enclosures and supporting equipment are accomplished and evaluated. Recommendations for improvements with respect to hardware and test and evaluation procedures are included in anticipation of ultimate use in U. S. Navy SCUBA operations.

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# I. INTRODUCTION

This thesis is the third in a series which addresses the problem of communications with the teathered diver. Lt. John Hamilton [Ref. 1] designed the Diver Monitoring System (D.M.S.) which transmits ultrasonic pulses indicating the diver's heart rate. Surface personnel can thus monitor the diver's general physical state. A C.W. feature was also incorporated which permits limited communication with the surface. While the requirement for a physiological monitoring system is certainly valid from a biomedical standpoint, voice communication is a more desirable and effective means of conveying information from diver to diver or from ship to diver. Lt. David Steere accomplished the design and construction of a frequency modulated communications system for free swimming divers [Ref. 2]. He made great progress in identifying peripheral equipment, conceptualizing the electronics package and investigating the theoretical aspects associated with DUCS I (Diver Underwater Communications System). DUCS II, as described herein, defines improvements or alternative methods concerning the electronic circuitry, underwater enclosures and other supporting equipment. This frequency modulated system, DUCS II, was constructed and evaluated, and recommendations for improvements are made in anticipation of ultimate use in U. S. Navy SCUBA operations.

## II. MODULATION CONSIDERATIONS

Frequency modulation of an acoustic carrier is the mode of transmission selected for use in DUCS II. Since the concepts of frequency modulation are available in numerous texts, considerations herein will be limited to a brief review of the salient aspects of F. M. and applications with respect to DUCS II.

DUCS II operates in narrow band F. M. The value of beta (maximum phase angle deviation is low, less than one for the high end of the audio band. Considering a single tone modulating signal, the smallest acceptable bandwidth is given by: bandwidth = $2(\beta + 1)f_m$, where $f_m$ is the frequency of the modulating signal. Assuming a $f_m$ of 3 KHz and receiver band-pass of 10 KHz centered on the carrier, beta is thus defined at approximately 0.7. The trade-off with beta is clear: high beta means larger bandwidth required.

DUCS II incorporates use of a mixer to translate the 40 KHz carrier to 455 KHz for the purpose of enhancing stability by splitting the R. F. and I. F. operating frequencies. This mixing process has no effect on the frequency deviation, where frequency deviation, $\Delta f$, is the change in frequency either side of carrier during modulation. (Beta is alternatively defined as the ratio of the maximum frequency deviation, $\Delta f$, to the modulating frequency, $f_m$.) Band-pass in DUCS II for the R. F. filter and I. F. strip are both set at 10 KHz. The ratio detector has a

frequency response of 20 KHz centered at 455 KHz.

The DUCS II receiver as tested provides good quality audio with an audio frequency range of 100 Hz to 4 KHz. Operation in the frequency modulation mode is considered satisfactory.

# III. ELECTRONICS SYSTEM DESCRIPTION

## A. SYSTEM OVERVIEW

A block diagram representation of DUCS II is depicted in figure 1. All electronics are contained in a single watertight enclosure through which connections are made for microphone, earphones and hand-held push-to-talk switch. Components are mounted on two separate boards. One carries the single conversion superhetrodyne-type receiver; the other houses the transmitter, voltage regulator and solid-state transmit/receive switch. Prototype design flexibility was afforded by using a wire-wrap technique. DIP sockets are used to mount both integrated circuits and most capacitors, resistors and diodes. Fourteen rechargable NICAD cells provide symmetrical positive and negative supply voltages, common to both transmitter and receiver boards.

The DUCS II transmitter remains essentially as conceived by Lt. Steere [Ref. 2]. However, no continuous wave mode is incorporated and the pseudo-logarithmic amplifier used to filter VCO output has been eliminated. The inherent bandpass characteristics of the transducer and impedance matching inductor are sufficient to filter the harmonics of the triangle waveform generated by the voltage controlled frequency modulation generator.

Power is applied to the receiver and all but the final amplifier sections of the transmitter by positioning an external magnet which

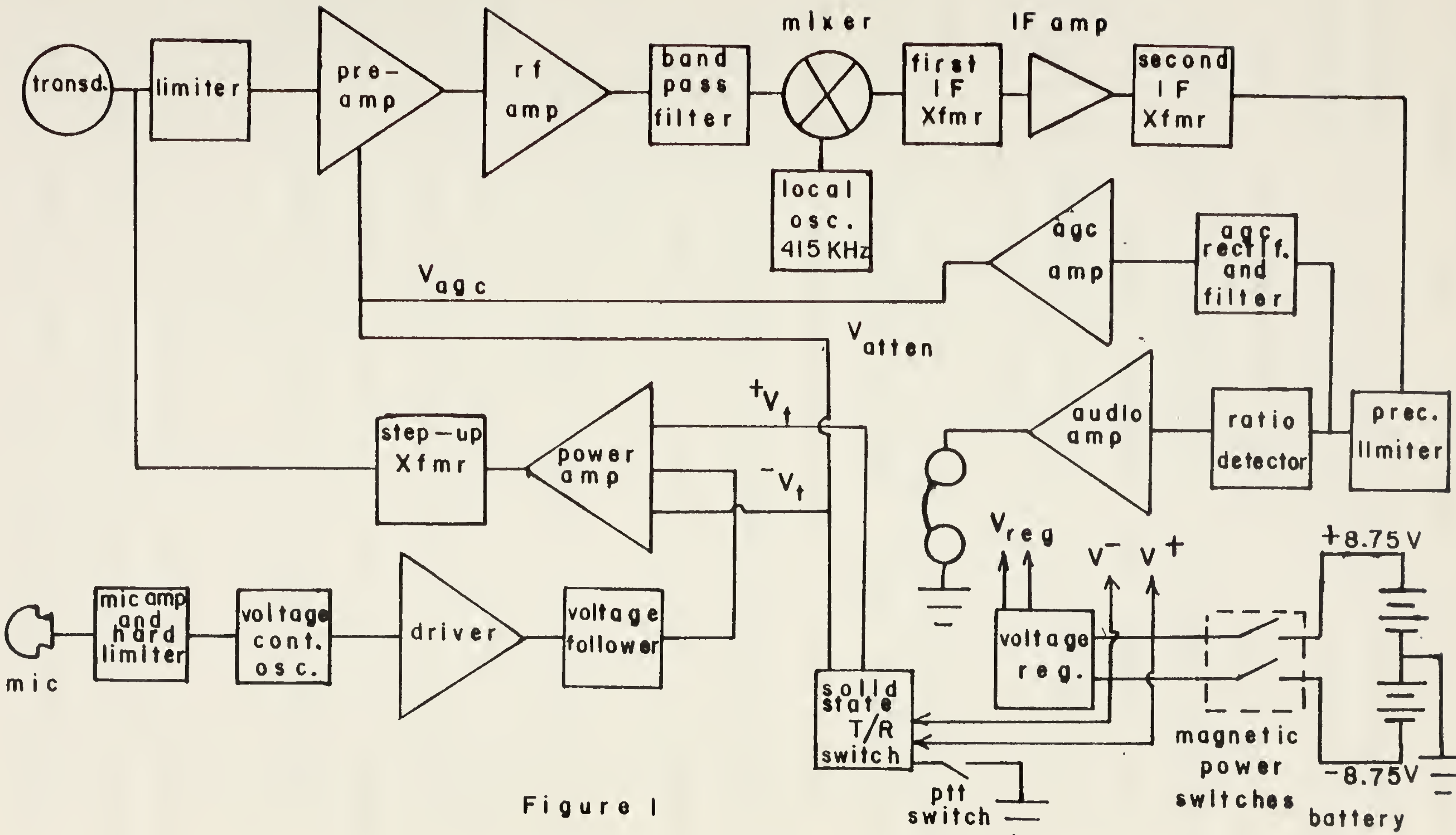



Figure 1

Electronics System Block Diagram

closes internal power switches. Audio voltages are generated at the microphone located in the full face mask which then pass through a precision limiter set to preclude over-deviation. A voltage controlled oscillator then translates these audio voltages to a frequency modulated signal. Final transmitter amplification is accomplished when the diver's hand-held push-to-talk switch is depressed, which applies driving voltages to the transmitter power amplifier. A set-up transformer passes the final signal to the impedance matching network and ultimately to the transducer for transmission. Transmitter design provides for frequency deviation adjustment up to 10KHz. Voltage controlled oscillator center frequency and microphone gain may be selected within limits. Keying of the transmitter simultaneously attenuates the receiver front end, allowing only sufficient gain for the diver to monitor his outgoing transmissions.

The receiver is always energized. Under no key conditions, the only attenuation is provided by AGC circuitry. Communications coming in from the transducer pass through a simple diode limiter to a pre-amplifier where attenuation access is afforded for both automatic gain control and transmit key attenuation. After the signal passes through a high gain R. F. amplifier, an active band-pass filter then rejects noise outside the desired frequency range. This well processed signal is then mixed with a local oscillator output operating at 415 KHz to produce a sum frequency at 455 KHz. The difference frequency,

other harmonics and noise are attenuated in a highly selective I. F. strip composed of two double-tuned I. F. transformers separated by an I. F. amplifier. A precision limiter then removes amplitude variations and generates the voltages from which A. G. C. is eventually derived. F. M. detection is achieved using a ratio detector. Finally, a single integrated circuit audio amplifier with volume control drives low impedance dynamic headphones. There is no squelch circuitry, although add-on at a later time is envisioned.

A highly stable voltage regulator furnishes a precision reference voltage for both the receiver local oscillator and the transmitter V.C.O.

## B. DESCRIPTION OF CIRCUITRY

The accompanying circuit diagrams are intended to provide a basic description of DUCS II circuitry. Table I is a complete listing of semiconductor devices as well as miniature coil components. The transmitter circuit board as installed is shown in figure 2. Figure 2 is a photograph of the receiver circuit board. Detailed circuit description first examines the supporting electronics, the voltage regulator and solid state transmit/receive switch. Transmitter and receiver operation are then considered.

### 1. Voltage Regulator

The transmitter V. C. O. and local oscillator in the receiver are Signetics 566 function generator integrated circuits which have a

TABLE I

LISTING OF MAJOR ELECTRONICS COMPONENTS

I. TRANSISTORS

| | | | |
|---|---|---|---|
| Q1 | SK3512 | Q8 | SK3513 |
| Q2 | 2N3706 | Q9 | 3N187 |
| Q3 | 2N3486 | Q10 | 3N187 |
| Q4 | 2N3866 | Q11 | 3N187 |
| Q5 | 2N3706 | Q12 | 3N187 |
| Q6 | SK3512 | Q13 | 3N187 |
| Q7 | 2N3702 | | |

II. DIODES

| | | | |
|---|---|---|---|
| D1-D4 | 1N277 | D13 | 1N64F |
| D5-D6 | 1N64F | D14-D15 | 1N277 |
| D7-D10 | 1N277 | ZD1-ZD2 | 1N750 |
| D11-D12 | 1N277 | ZD3-ZD4 | 1N750 |

III. INTEGRATED CIRCUITS

| | | |
|---|---|---|
| U1 | SE550 | VOLTAGE REGULATOR |
| U2 | LM741 | OP AMP |
| U3 | SE566 | FUNCTION GENERATOR |
| U4A | LM747 | DUAL OP AMP |
| U4B | LM747 | DUAL OP AMP |
| U5 | BB3508J | OP AMP |
| U6 | LM741 | OP AMP |
| U7 | SE566 | FUNCTION GENERATOR |

| | | | |
|---|---|---|---|
| U8 | BB3508J | OP AMP | |
| U9 | LM380N-14 | AUDIO AMP | |

IV. INDUCTORS AND TRANSFORMERS

T1 TRANSMITTER STEP-UP TRANSFORMER:
17 TURNS #18 AWG (P)
177 TURNS #23 AWG (S)
AMIDON T-80-2 TOROID

| | | | |
|---|---|---|---|
| T2 | VARIABLE INDUCTOR | MILLER P.N. 9058 | .65-1.3mH |
| T3 | I. F. TRANSFORMER | MILLER P.N. 8807 | |
| T4 | I. F. TRANSFORMER | MILLER P.N. 8807 | |
| T5 | RATIO DETECTOR | MILLER P.N. 8805 | |

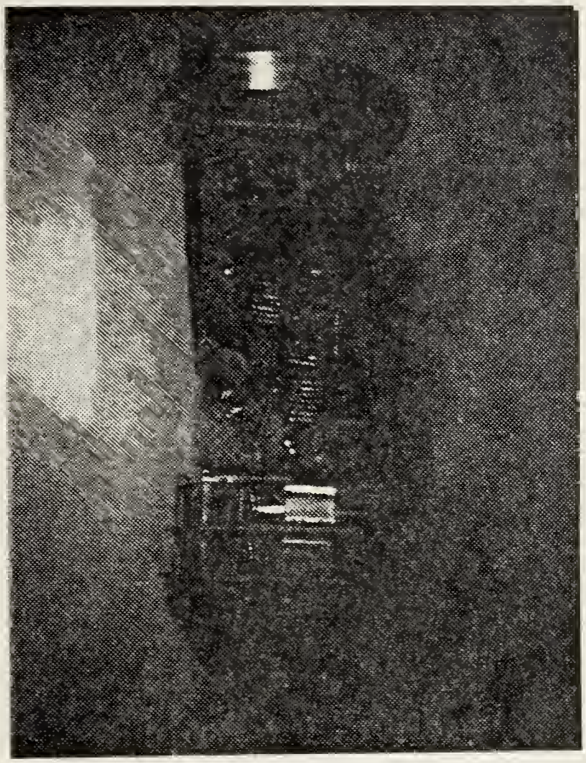


Figure 2

Transmitter

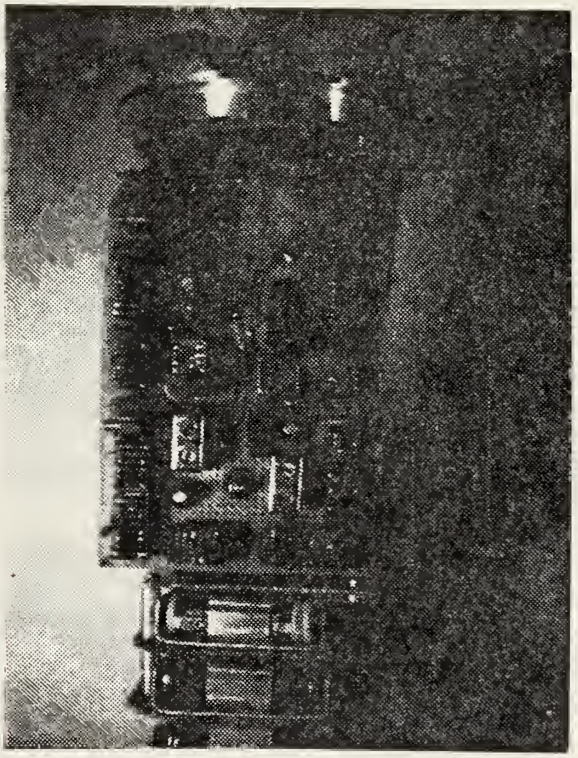


Figure 3

Receiver

published frequency drift in percent operating frequency of 1 to 2 percent per volt supply [Ref. 4]. A power supply variation of 1 to 2 volts is possible while still providing useful service. This sort of variation would generate frequency drifts of 1 KHz in the transmitter V.C.O. and 8 KHz in the local oscillator, putting the mixed frequencies well outside the I. F. bandpass. The voltage control circuitry of DUCS II is shown in figure 4. The Signetics SE 550 integrated circuit provides load regulation of within a few millivolts over several volts change in the supply. Q1 is a pass NPN transistor with a 2 ampere capability. $R_3$ is a potentiometer which controls $V_{reg}$; it is set at 10.00 volts with respect to the negative side of the battery power supply. The V.C.O. and local oscillator both require a minimum supply of ten volts and are similarly referenced. Complete short circuit protection is a design feature of the voltage regulator.

2. Solid State Transmit/Receive Switch

A solid state switch gates driving symmetrical positive and negative voltages to the final amplifier stage of the transmitter. The design is as utilized by Lt. Steere in DUCS I [Ref. 2], and is shown in figure 5. The push-to-talk switch grounds the base of Q2 when depressed. Q2 is normally off because base and emitter are at supply potential of -8.75 volts. With no current through Q2, there is no drop across $R_3$ and $R_7$ and the power supply voltages are on the collector and emitter of Q2. Depressing the P. T. T. switch forward

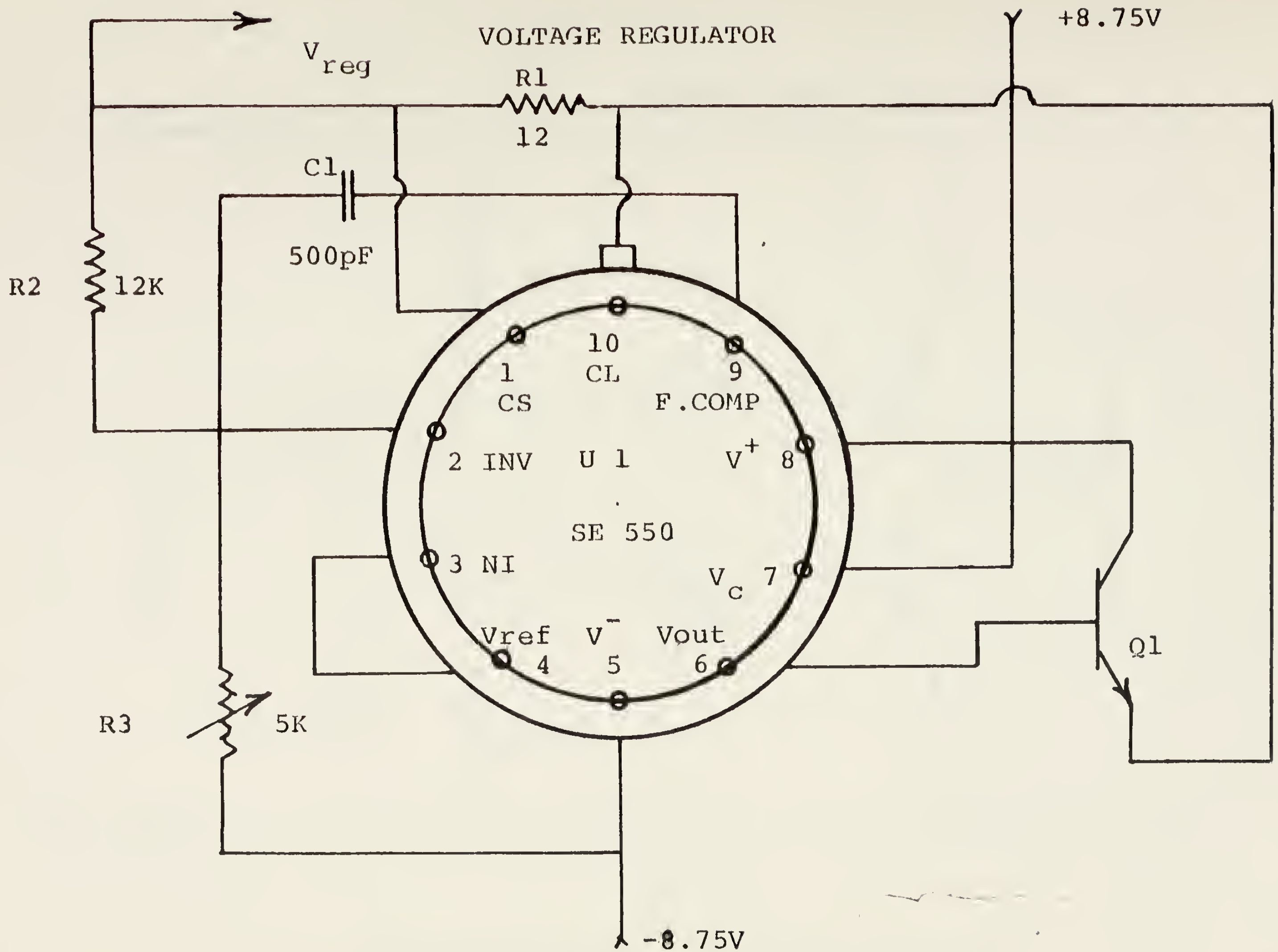



Figure 4

Schematic Diagram: Voltage Regulator

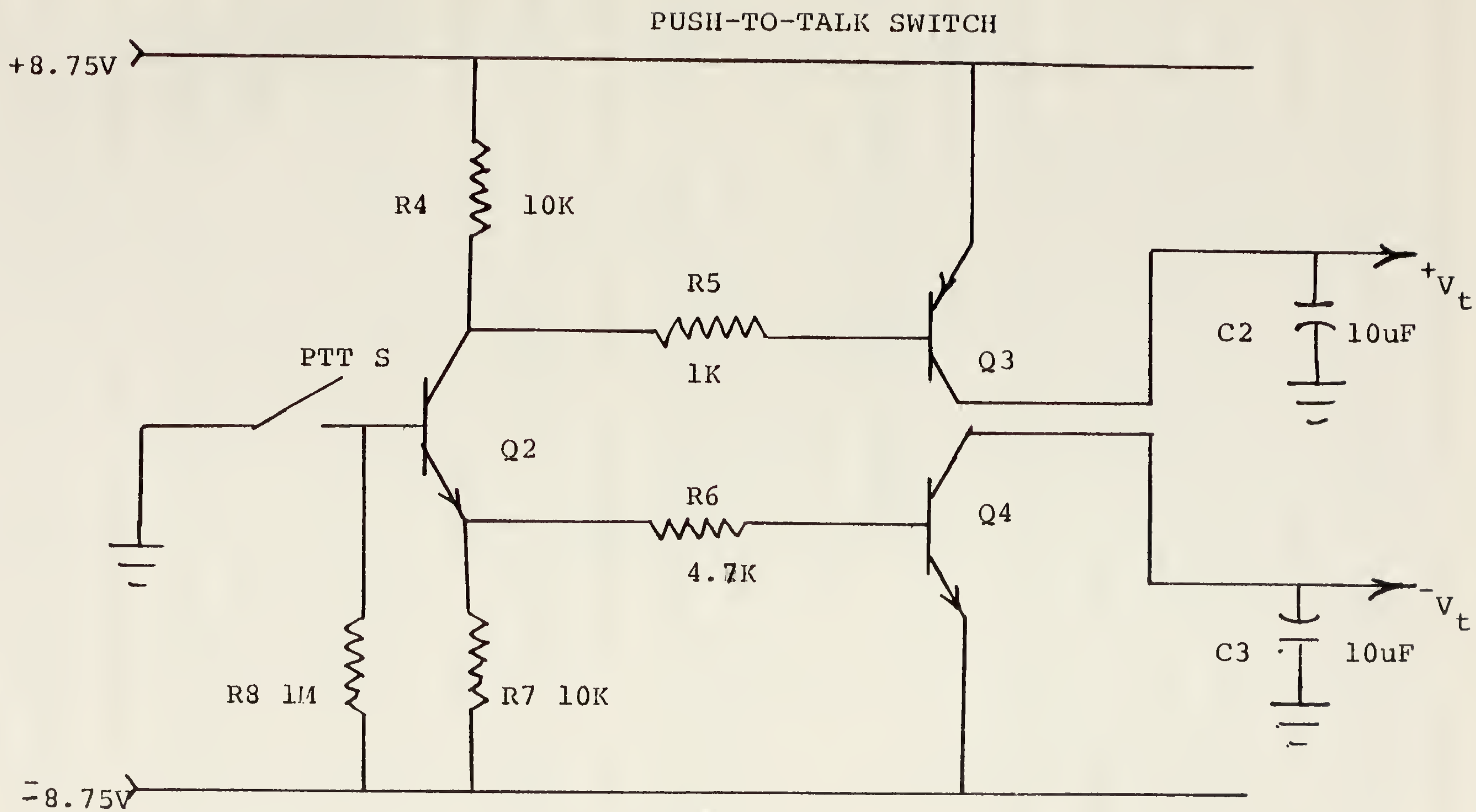



Figure 5

Schematic Diagram: Solid State T/R Switch

biases the base/emitter junction of Q2. With current now flowing through Q2, the voltage drop across $R_4$ and $R_6$ drives the bases of $Q_3$ and $Q_4$ toward ground. $Q_3$ and $Q_4$ are therefore driven on, passing the transmitter voltages, $\pm V_t$. The negative voltage, $-V_t$, is additionally used as the source for the attenuating voltage applied at the front end of the receiver during transmitter keying. The pass transistors $Q_3$ and $Q_4$ were specifically selected for their low resistance values when saturated, thus maximizing the driving voltages presented to the transmitter power amplifier.

3. Detailed Transmitter Analysis

a. Microphone Amplifier/Limiter

Audio voltages generated in the ceramic microphone are capacitively coupled into a microphone amplifier/precision bridge limiter, the design of which is published by the Burr-Brown Research Corporation [Ref. 3]. See figure 6. Temperature sensitivity and non-linearity of the diodes in the bridge are reduced using the high gain of an operational amplifier (U2). This circuit limits both the positive and negative excursions and produces sharp clipping at the design voltages. The forward diode voltage drop and non-linearity are reduced by an amount equal to the open loop gain of the op amp; this is because the diode bridge is contained within the feedback loop. Figure 7 is a photograph of the limiting as accomplished by the circuitry when driven by a 10 KHz sinusoidal signal. A graphical indication of

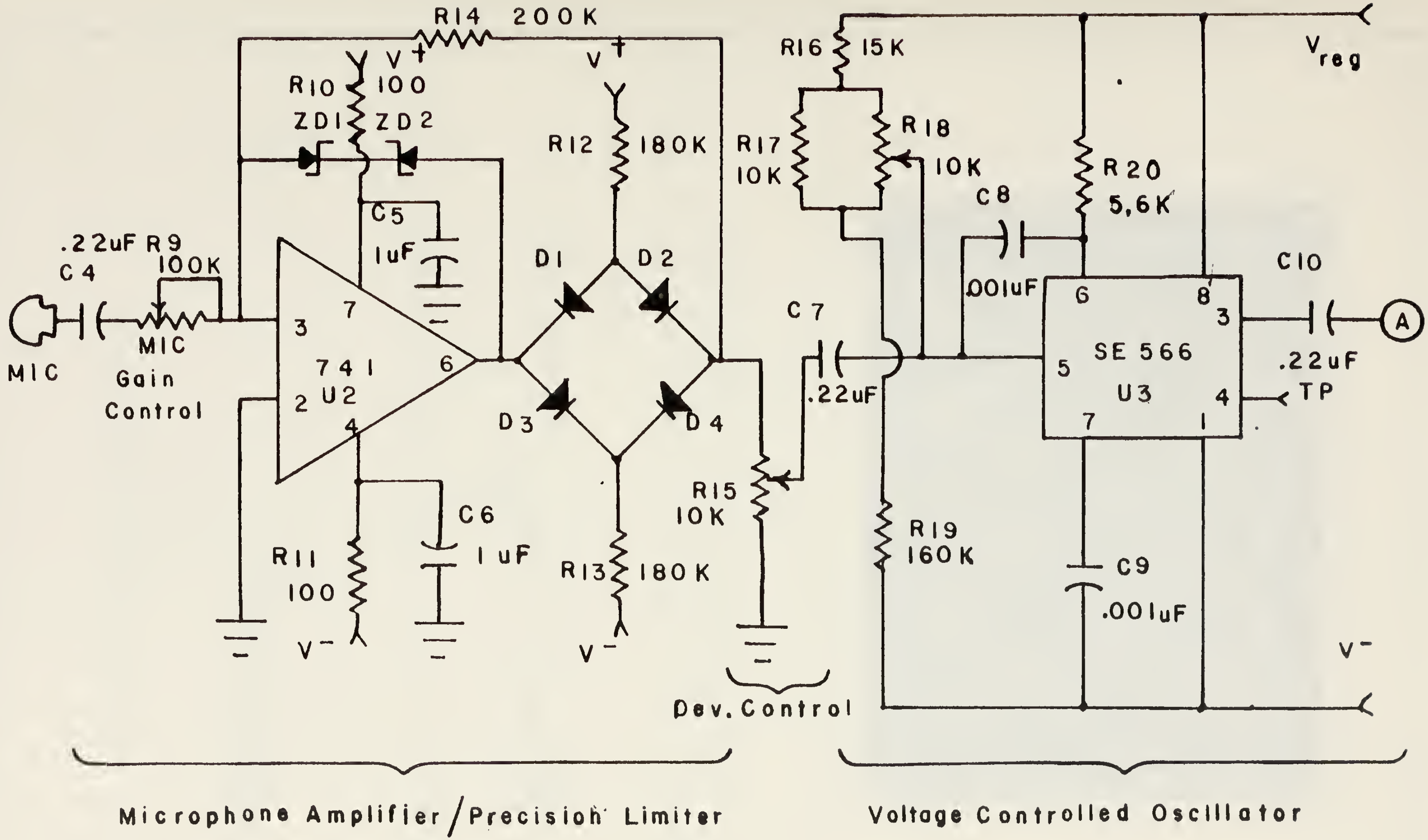



**Figure 6**

Schematic Diagram: Transmitter Microphone Amplifier/Limiter and V.C.O.

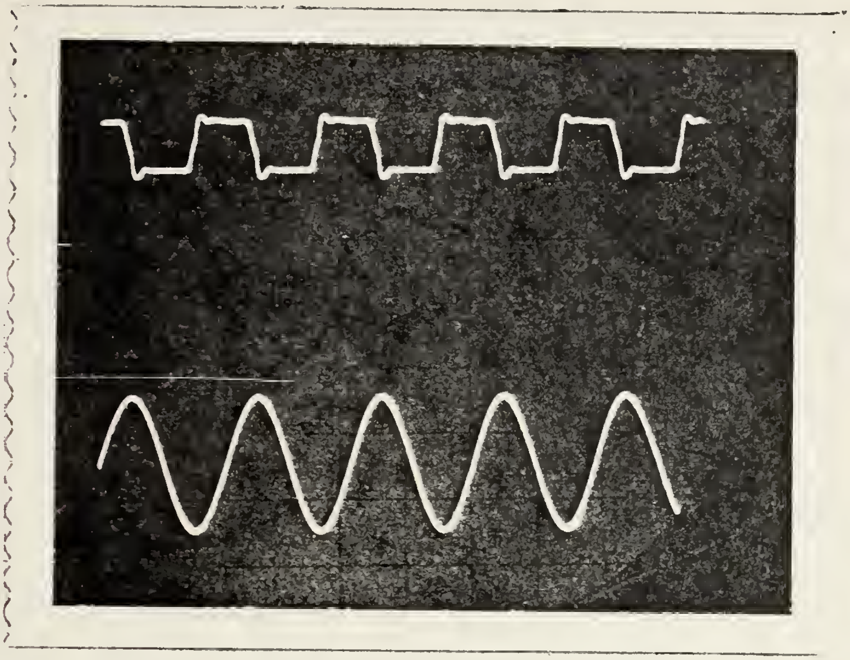


Figure 7

Photograph of Transmitter Limiting Action

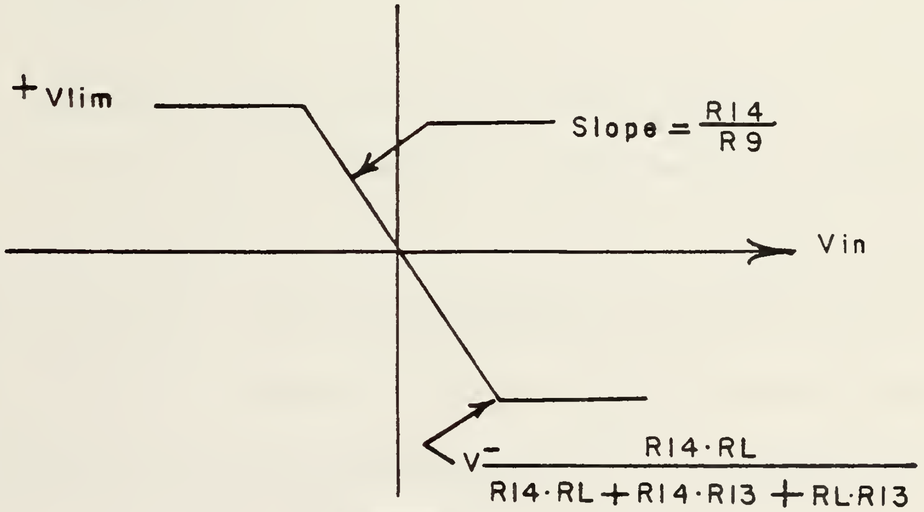



Figure 8

Graphic Representation of Limiter

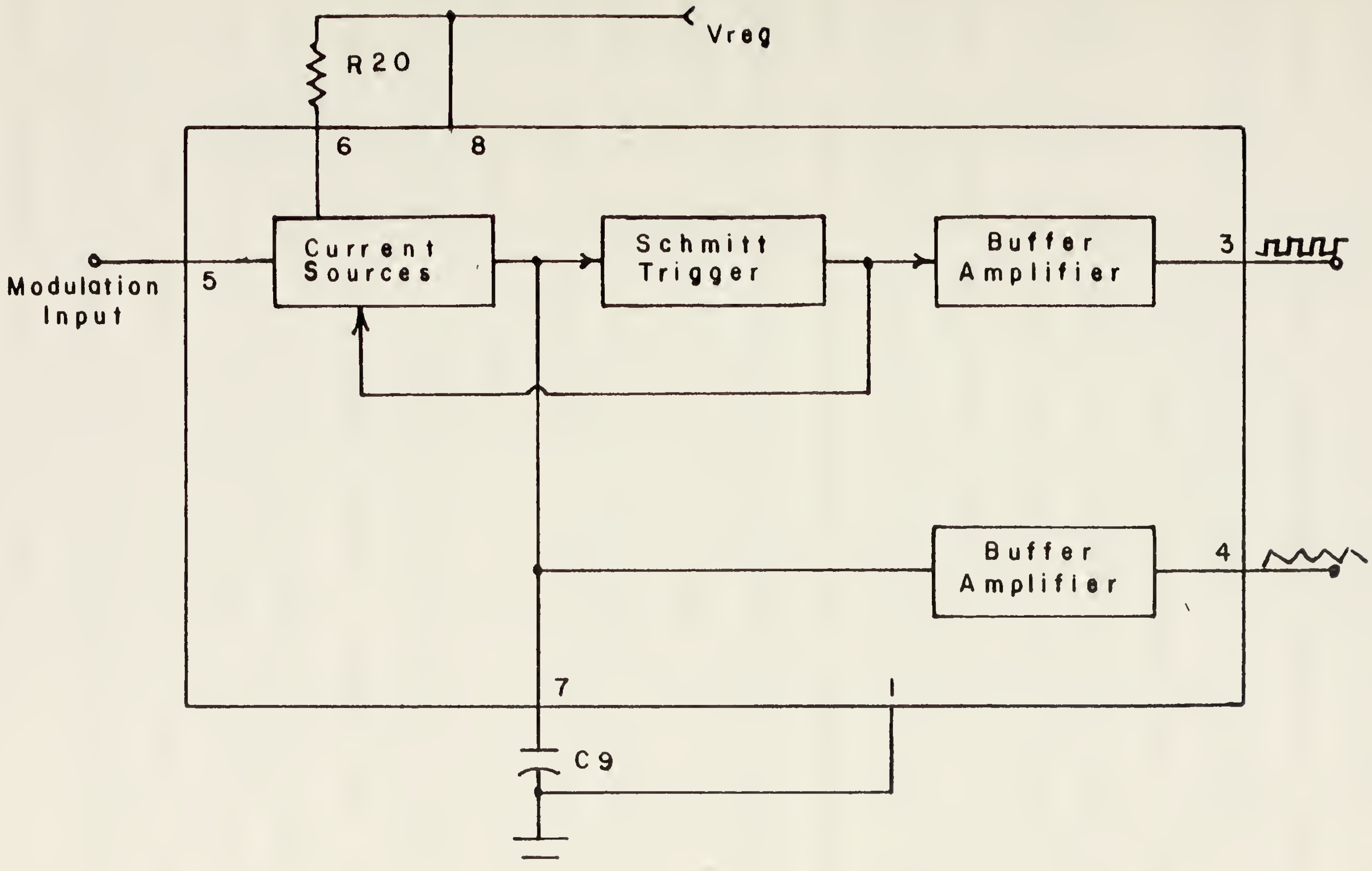
the clipping action is presented in figure 8. This circuit is also utilized in the receiver. A brief description of the design parameters is included here. The slope of the graph is determined by the ratio of the feedback resistance, R14, to the input resistor at the op amp, R9. The clipping action is to be accomplished under the loudest bubble noise or loud voice conditions, and is designed for 100 millivolts. A satisfactory approximation is given by:

$$\pm V_{lim} = \pm \frac{R_{14}R_L}{R_{14}R_L + R_{14}R_{12} + R_L R_{12}} V_s$$

With $V_{lim}$ at .1 volts $R_{14} = 200K$, $V_s = \pm 8.75$ volts and $R_L$ at 5K, the symmetrical resistors $R_{12} = R_{13}$ are calculated at 422K ohms. Standard values of 390K ohms are used. It is noted that $R_l$ is the impedance seen looking into the V.C.O./F.M. generation section of the transmitter and is taken to be about 5K ohms.

b. F.M. Generation/Voltage Controlled Oscillator

The heart of the transmitter is contained in the voltage controlled oscillator section which converts the voice-generated signals into frequency modulation. Excellent realization is found in Signetics Corporation's SE/NE 566 monolithic integrated circuit function generator [Ref. 4]. A block diagram representation of this F.M. modulator is given in figure 9. The free-running frequency is determined by an external resistor and capacitor or by the current injected into pin 6. Modulation about this center frequency is driven by the voltage presented




NE\SE566

Figure 9

Voltage Controlled Oscillator Block Diagram

at the input, pin 5. A typical published figure for frequency stability is 100 p.p.m. per degree centigrade. Assuming a worst case situation of ambient water temperatures ranging from -1°C to +27°C, stability equates to a frequency drift on the order of a hundred Hz, which is acceptable. The limiting consideration is the receiver passive band-pass characteristic of about 10 KHz centered on 455 KHz.

Transmitter center frequency and local oscillator frequency in the receiving unit must be tightly held to design specifications or audio distortion will be generated in passage through the I. F. strip. Minimum supply voltage for the function generator is 10 volts; therefore the circuit is referenced to V- vice system ground. Although typical frequency drift is one percent of free running frequency per volt change in supply voltage, the use of the SE 550 voltage regulator effectively precludes adverse effects of battery fluctuations. Prototype transmitters were observed to be very stable in frequency under changes of several volts in supply voltage. Output is taken at pin 3, the square wave, while the other output at pin 4 is used as a frequency test point. The buffer amplifiers at these terminals minimize the effect of external loading on the center frequency stability. Unlike DUCS I, special provision is not made to filter or convert the square wave output to a sinusoidal waveform. The square wave is rich in fundamental frequency; filtering is inherent in the transducer resonant characteristics. Frequency deviation is determined by the magnitude of the applied voltage which is adjustable using the potentiometer R15.

c. Transmitter Final Amplifier Stage

A single integrated circuit (Ur), with two operational amplifiers contained within, is utilized as the driver and voltage follower in the transmitter. See Figure 10. Driver gain is approximately determined by the ratio of the feedback resistor $R_{24}$ to that of the input resistance $R_{23}$; a gain of about nine (19 dB). The second half of the I. C. is the voltage follower with the power amplifier contained in the feedback loop. According to Steere [Ref. 2], the voltage follower provices an input impedance greater than 400 MOhms with only a few ohms output impedance. The power amplifier is formed by two pairs of transistors arranged in complementary push-pull. The pass transistors Q6 and Q7 deliver power directly from the battery to the step-up transformer. The arrangement of the voltage follower (U4B) and the pass transistors reduces the non-uniformity by the multiplicative inverse of the amplifier gain. Higher current gain and lower output impedance are realized with the Darlington arrangement than with single power transistors alone.

d. Step-Up Transformer

As shown later in the underwater acoustics section, V, the transducer impedance at 40 KHz is nearly resistive at 520 Ohms. In view of the low output impedance of the power amplifier, transformer coupling is utilized. An Amidon T80-2 toroid core was wound with 17 turns of no. 18 wire as the primary. The secondary has 117 turns of

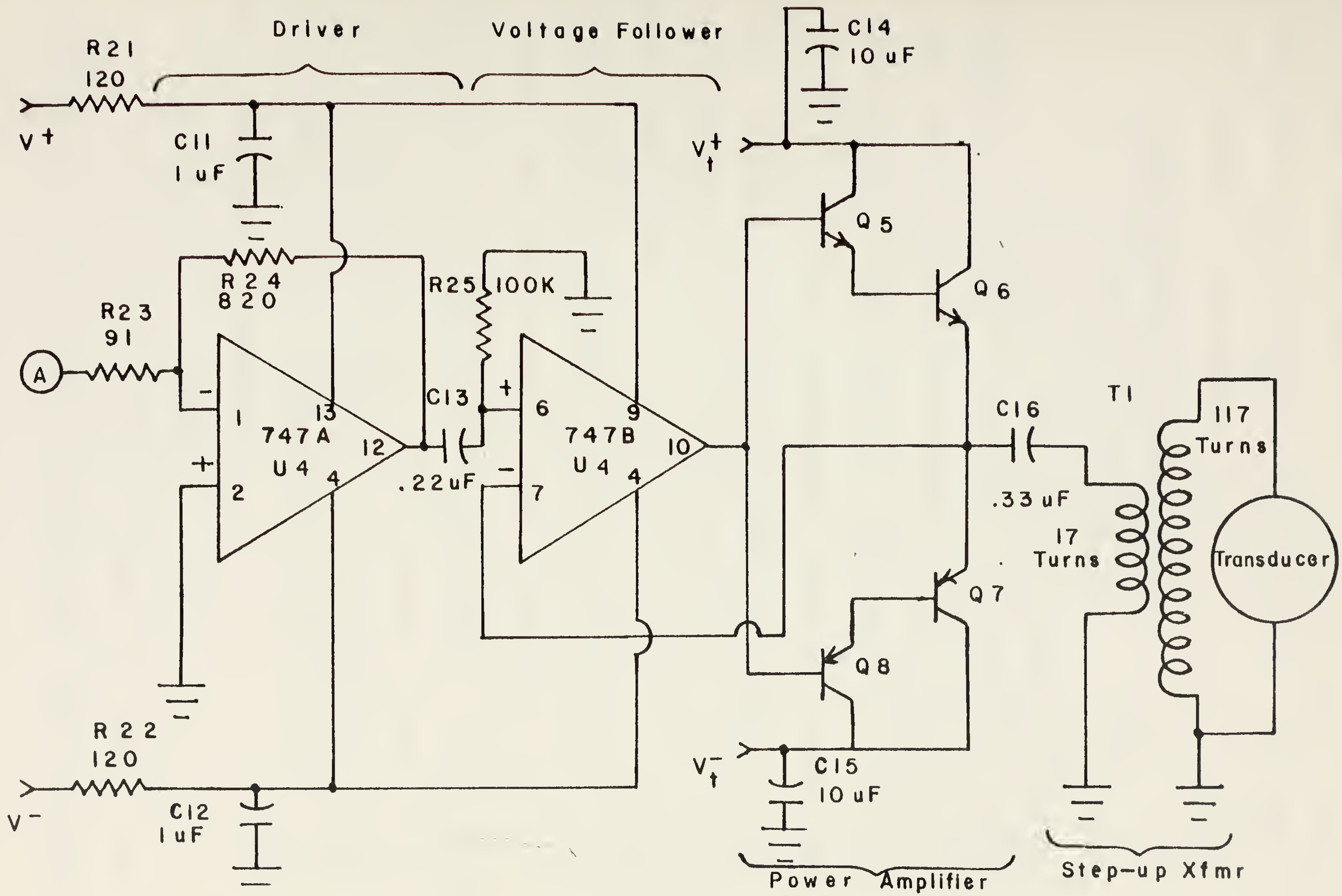



Figure 10

Schematic Diagram: Transmitter Driver, Voltage Follower and Power Amplifier

no. 23 wire, yielding a secondary to primary ratio of 6.9. The given turns ratio presents an impedance of about 11 Ohms. The voltage swing at the transducer is approximately 50 volts.

4. Detailed Receiver Analysis (Figure 11)

a. Transmit/Receive Switching

The single transducer element is employed in both the transmit and receive functions as follows: C17 at 40 KHz incoming signal generates approximately 8 Kohms of capacitive reaactance to current inbound from the transducer. Diodes D5 and D6 will not conduct until about .4 Volts, presenting high impedance at normal signal levels. Under no key conditions, the transmitter is disabled. The power Darlington transistors in the non-conducting condition present high impedance to the step-up transformer primary. The secondary of the transformer may then be considered at infinite impedance. By voltage divider action, virtually all of the incoming signal appears at gate number 2 of the MOSFET receiver pre-amplifier. Under transmit conditions, a negative voltage of .6 volts is applied to gate 1 of the pre-amp. By pinching off the MOSFET, 12 dB of attenuation is realized. This is within the A.G.C. capability of the receiver and permits the diver to monitor his own transmissions.

b. Receiver Pre-Amplifier

All transistors in DUCS II receiver are of the dual gate, n-channel, metal oxide field effect transistor type. They are connected

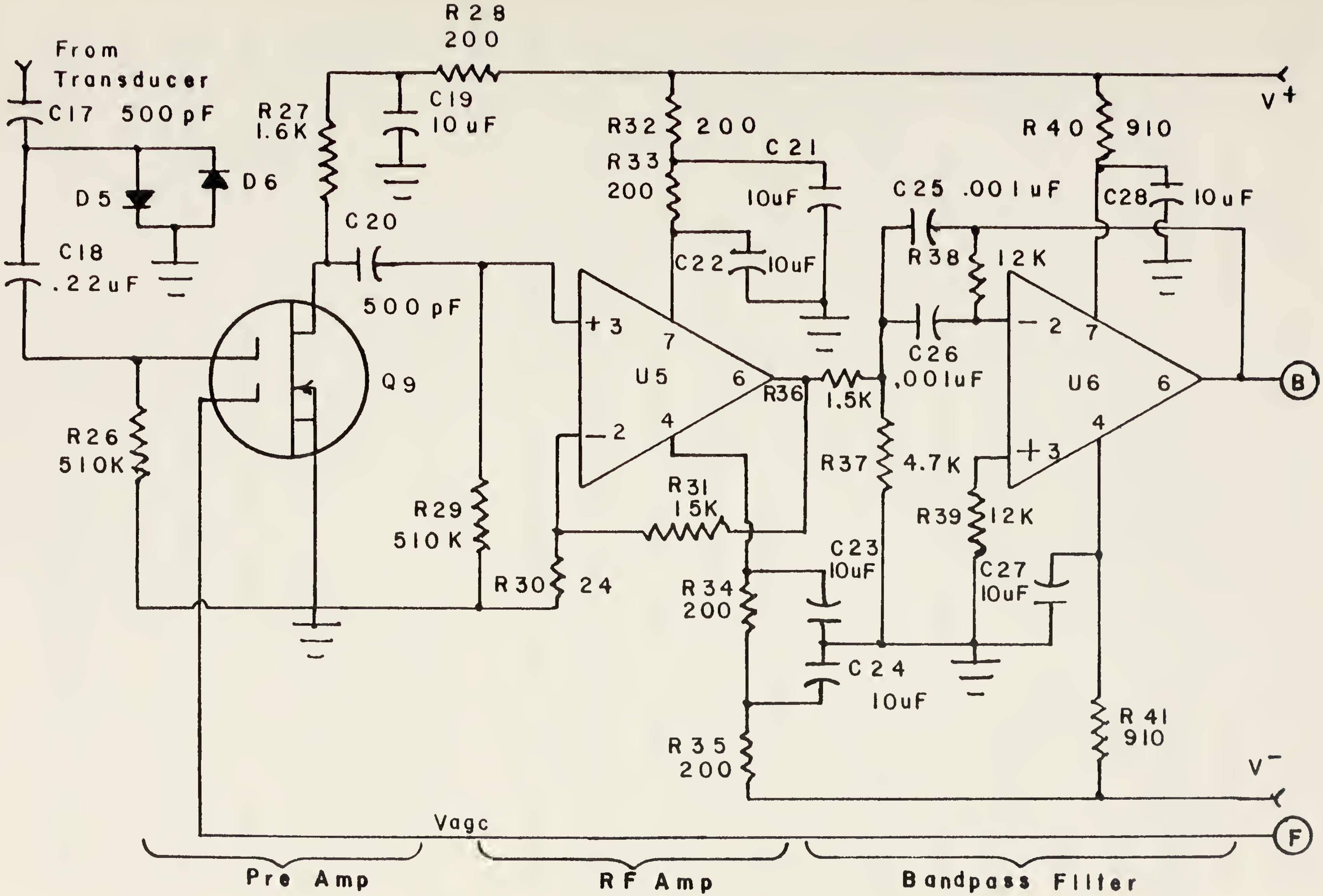


Figure 11

Schematic Diagram: Receiver Pre-Amp, R. F. Amp and Band-Pass Filter

in triode or tetrode as the situation warrants. The advantages of MOSFETs are many. They exhibit extremely high input resistance due to the insulated gate construction of the device. The high input resistance allows the use of simple electron tube-type biasing. Large positive and negative input excursions are possible without the diode current loading of bipolar transistors. This means greater dynamic ranges than JFET or bipolar transistors and superior application as A.G.C. devices. Finally, the MOSFET is a low noise unit with two gates optional, one for the signal and one for A.G.C. access, making it ideal for use as a receiver first stage amplifier. The dual gate MOSFET may be considered as two single-gated MOSFETs connected in cascode with channels in series and separate inputs. Output current is therefore subject to the control of either gate. In DUCS II, the A.G.C. signal and transmit key signal for front end attenuation are applied to gate 1, while the incoming signal is capacitively coupled into gate 2.

Pre-amplifier attenuation as a function of applied A.G.C. voltage is shown graphically in figure 12. The transistor amplifier (Q9) is connected in the common source configuration, the output of which drives the R.F. amplifier. Gain under no A.G.C. conditions is approximately 3.5 (10.9 dB). Power supply voltages are judiciously filtered by the R/C combination of R28 and C19. This practice, intended to damp unwanted oscillations, is continued throughout the receiver.

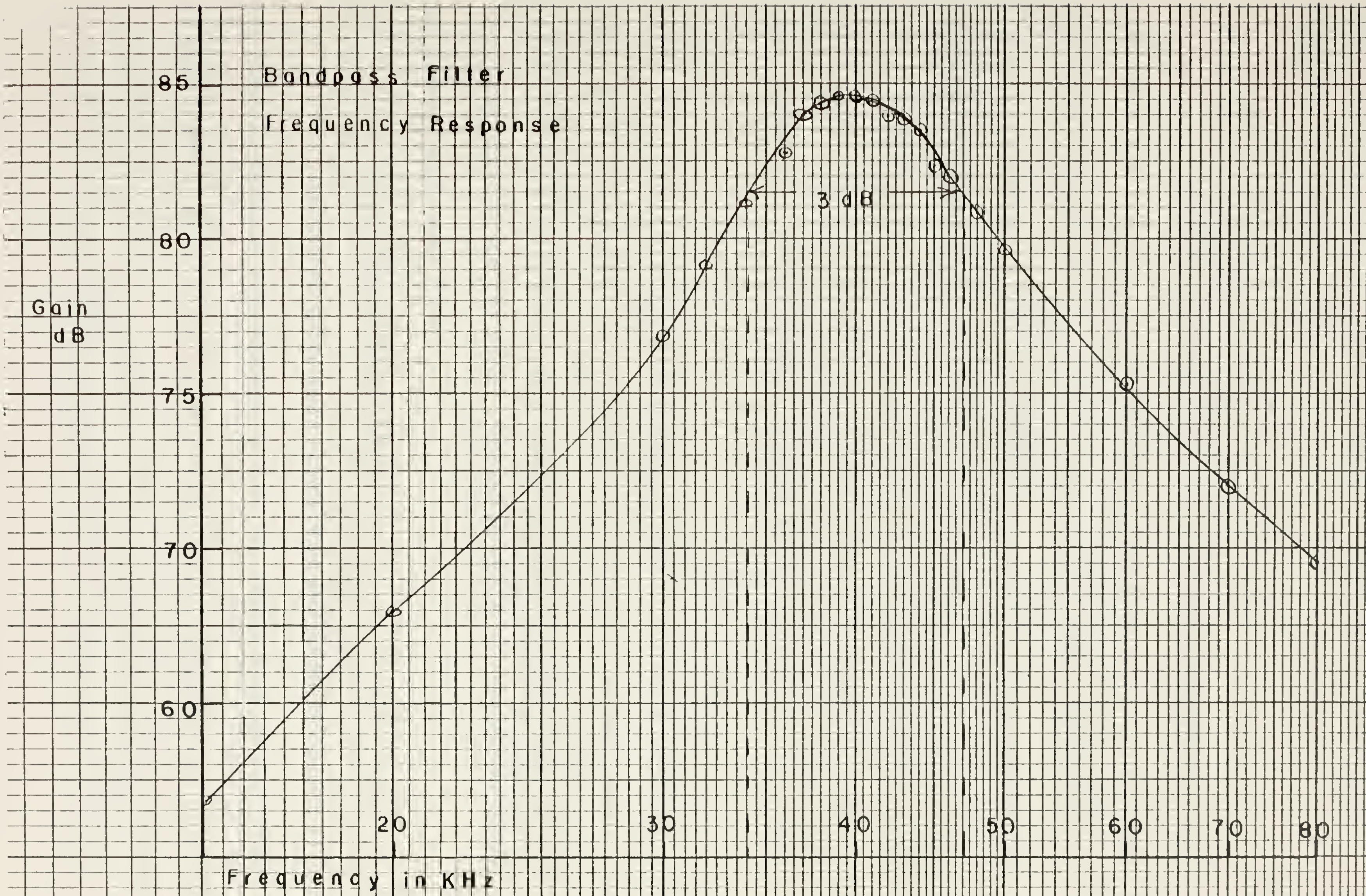



Figure 12

Receiver Band-Pass Frequency Response

Very high gain devices, such as the R.F. amplifier, are double filtered at the supply inputs.

c. Receiver High Gain R.F. Amplifier

Immediately following the pre-amp is a high gain R.F. amplifier (U5) which delivers an impressive gain of about 1355 (62.6 dB). Gain of this magnitude is not easily generated from a single OP The open loop gain of a 741 OP AMP, for example, has a unity gain bandwidth product of 1 MHz. The OP AMP selected for use as R.F. amp is a Burr-Brown 3508J which has a published U.G.B.P. of 100 MHz. Additionally, only 25 nA of bias current is required. Attendant input impedance is 300 MOhms shunted by three pF [Ref. 5]. The OP AMP is connected in non-inverting configuration as seen in figure 11. This arrangement allows for isolation of the feedback signal from the input signal, enhancing stability under high gain conditions. R29 provides a D.C. current path to ground for the non-inverting input of the OP AMP. Output of the R.F. amplifier drives the active band-pass filter.

d. Active Band-Pass Filter

An active band-pass filter was designed using the criteria set forth by Hilburn and Johnson in Ref. 6. Since high gain is not a requirement in this section, a standard LM 741 OP AMP (U6) proved to be adequate. Three resistors, R36, R37 and R38 and two capacitors C25 and C26 determine the parameters of the second order multiple

feedback band-pass filter. An excellent approximation is realized using the following design equations (C25=C26=C):

$$B = \text{bandwidth in radian frequency} = \frac{2}{R38\ C}$$

$$\omega_o^2 = \text{center radian frequency} = \frac{1}{R38\ C^2}\left(\frac{1}{R36} + \frac{1}{R37}\right)$$

$$\text{Gain (inverting)} = \frac{R38}{2\ R36}$$

Hilbrun and Johnson supplement the equations with graphical methods for determining values of capacitors and resistors. Adjustment to standard values available and to precise parameters desired is easily accomplished, noting the direct or inverse relationship of the above equations. The gain of this section of the receiver was observed to be 3.8 (11.6 dB) which compares favorably with the calculated gain of 4.0 (12 dB). Theoretical center frequency is 43 KHz versus the observed value of 40 KHz. The three dB points are at 34 KHz and 47 KHz which satisfies the band-pass requirements of the transmitted signal. A frequency response curve was plotted injecting a nominal 80 micro-volts at 40 KHz into the receiver pre-amplifier at gate two and measuring the response at pin six of the band-pass filter amplifier. The gain at 40 KHz is about 12,400 (82 dB) which substantiates the individual measured stage gains of 10, 62 and 11 dB at the pre-amp, R. F. amp and filter respectively. Figure 11 is a diagrammatic representation

of the receiver considered thus far, the pre-amp, R. F. amp and the band-pass filter. R39 serves to balance the impedances at the input gates to equalize currents and minimize offset voltages.

e. Receiver Intermediate Frequency Section (Figure 13)

The R. F. designated section of the receiver operates at the incoming frequency of 40 KHz with a five KHz F. M. deviation. This signal passes into an intermediate frequency section of the receiver which operates at 455 KHz. The first component of this I. F. strip is a mixer (Q 10) into which is fed a local oscillator sinusoid at 415 KHz and the received signal. The sum frequency at 455 KHz F. M. passes into the first I. F. transformer (T3), an I. F. amplifier (Q11) and a second double-tuned I. F. transformer where the difference frequencies generated in the mixing and other undesired noise are selectively attenuated. A limiter is the last item in the section operating at the I. F. before audio conversion is accomplished.

The mixer is a dual gate MOSFET connected in tetrode, common source arrangement with the received signal coupled into gate one and the local oscillator signal into gate two. Current output on the drain is directly passed to the input of T3, the first I. F. double-tuned transformer.

The local oscillator signal at 415 KHz is generated by a Signetics SE 566 function generator similar to the transmitter F. M. generator. The square wave presented at pin 3 is rich in fundamental

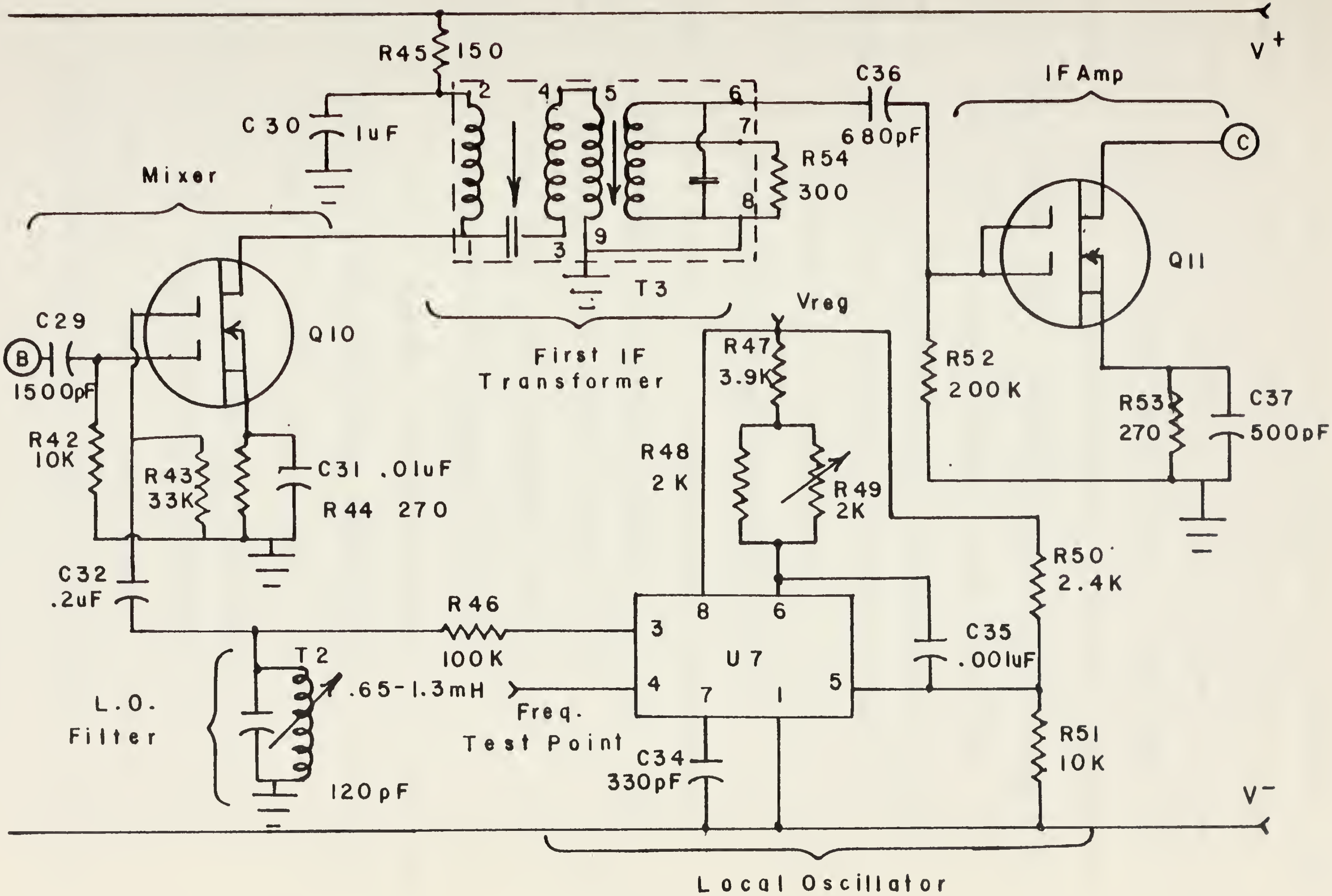



Figure 13

Schematic Diagram: Receiver Mixer, Local Oscillator, First I. F. Transformer and I. F. Amp

frequency; it is filtered by the parallel combination of T2 and C33. The waveform after filtering is a high quality sinusoid of about 1 volt peak to peak and is capacitively coupled into gate 2 of the mixer. R46 and R43 are used to drop the 4 volt peak to peak at the I. C. down to the 1 volt value at the mixer. A 1 microvolt signal at the receiver input appears as 12 millivolt at the mixer. Expected voltages at the R. F. gate of the mixer range then from tens of millivolts to about 200 millivolts. This was observed to be a satisfactory signal to oscillator ratio for good mixing. R49 allows for center frequency adjustment in the local oscillator to accommodate differences in individual integrated circuits.

The first I. F. transformer follows the mixer. It is a miniature double-tuned device (T3) manufactured by Bell Industries, J. W. Miller Division. Design center frequency is 455 KHz. R54 is a load resistor which sets the bandwidth at about ten KHz. Again, supply voltages are filtered by the R/C combination of R45 and C30. The second I. F. transformer (T4) is identical to the first.

The I. F. amplifier (Q11) buffers the two I. F. transformers and is connected in triode arrangement, common source. Initial intention was to use gate 2 as another point to which A. G. C. may be applied. Sufficient A. G. C. control is afforded in the pre-amplifier section on the receiver. The second gate is therefore shorted to gate 1. The I. F. transformers are sensitive to loading making the high impedance of the MOSFET particularly useful. Mixer, local oscillator, first I. F.

transformer and I. F. amplifier are shown in figure 13, while the second I. F. transformer and limiter are shown in figure 14. The frequency response of the I. F. strip is depicted graphically in figure 15. This curve is representative of the author's receiver as constructed and tuned. Referring to figure 15, the response peaks at 455 KHz with 3 dB points located approximately 5 KHz either side of center frequency. The curve is generated by injecting a nominal 100 microvolt signal at 455 KHz (no F. M. present) at gate 1 of the mixer. Output is measured at pin 6 of the limiter OP AMP (U8) where a gain at center frequency of 7 (17 dB) is observed.

Amplitude excursions in excess of 200 millivolts are removed in the precision limiter shown in figure 14. Again, a non-inverting configuration is used with signal into the non-inverting pin 3 while feedback is separately input to gate 2. Design criteria were previously considered in the transmitter section where limiting is used to preclude over-deviation. In the receiver, however, output is coupled into a buffer MOSFET amplifier with high input resistance. R65 is required to provide a load resistance for the limiting circuitry. 1000 ohms as $R_L$ limits at approximately 200 millivolts. This value is easily changed by varying R65. Components operating at the I. F. frequency, then, include mixer, I. F. transformers, I. F. amplifier and limiter. At this point the signal is ready for demodulation.

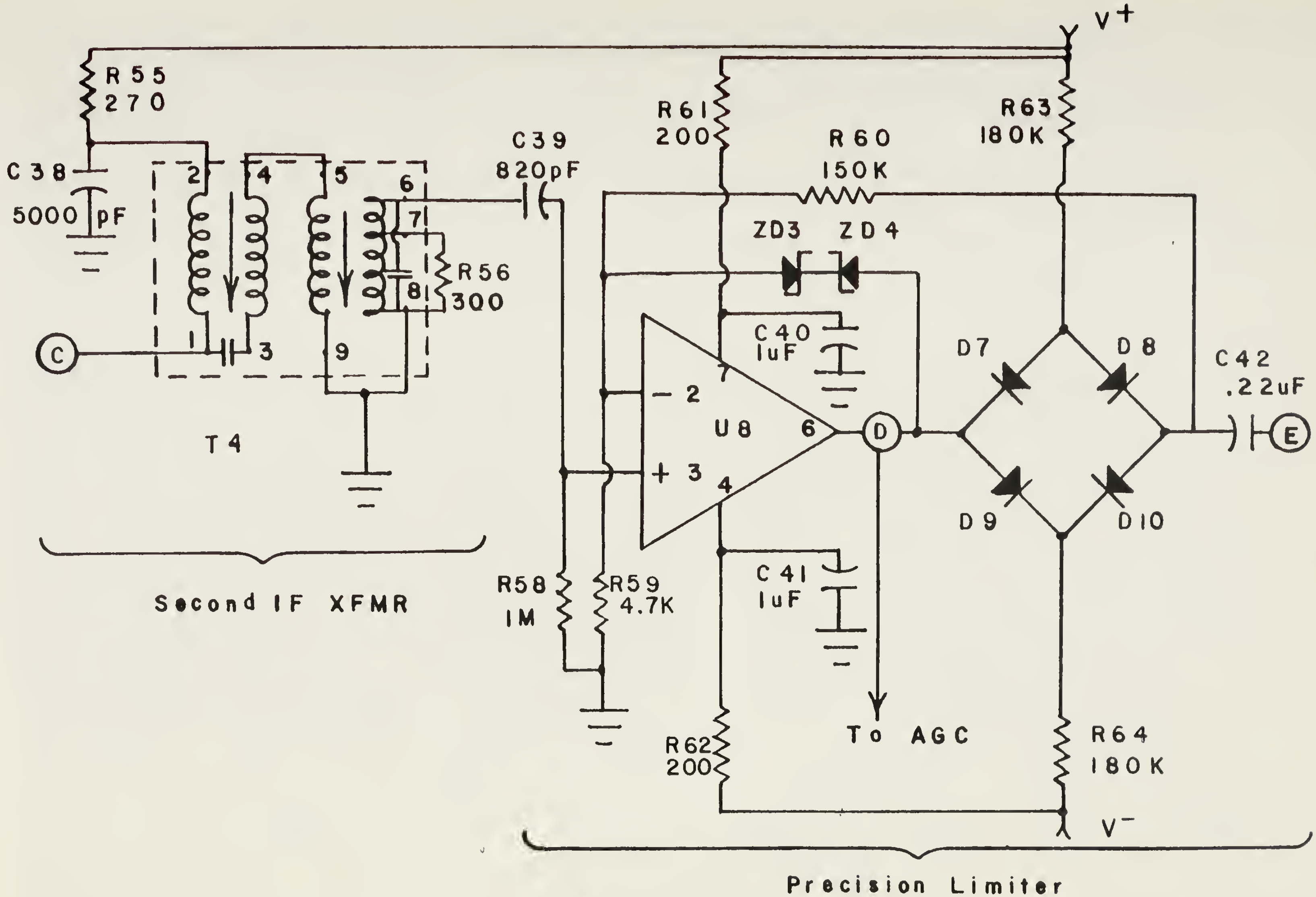



Figure 14

Schematic Diagram: Second I. F. Transformer and Precision Limiter

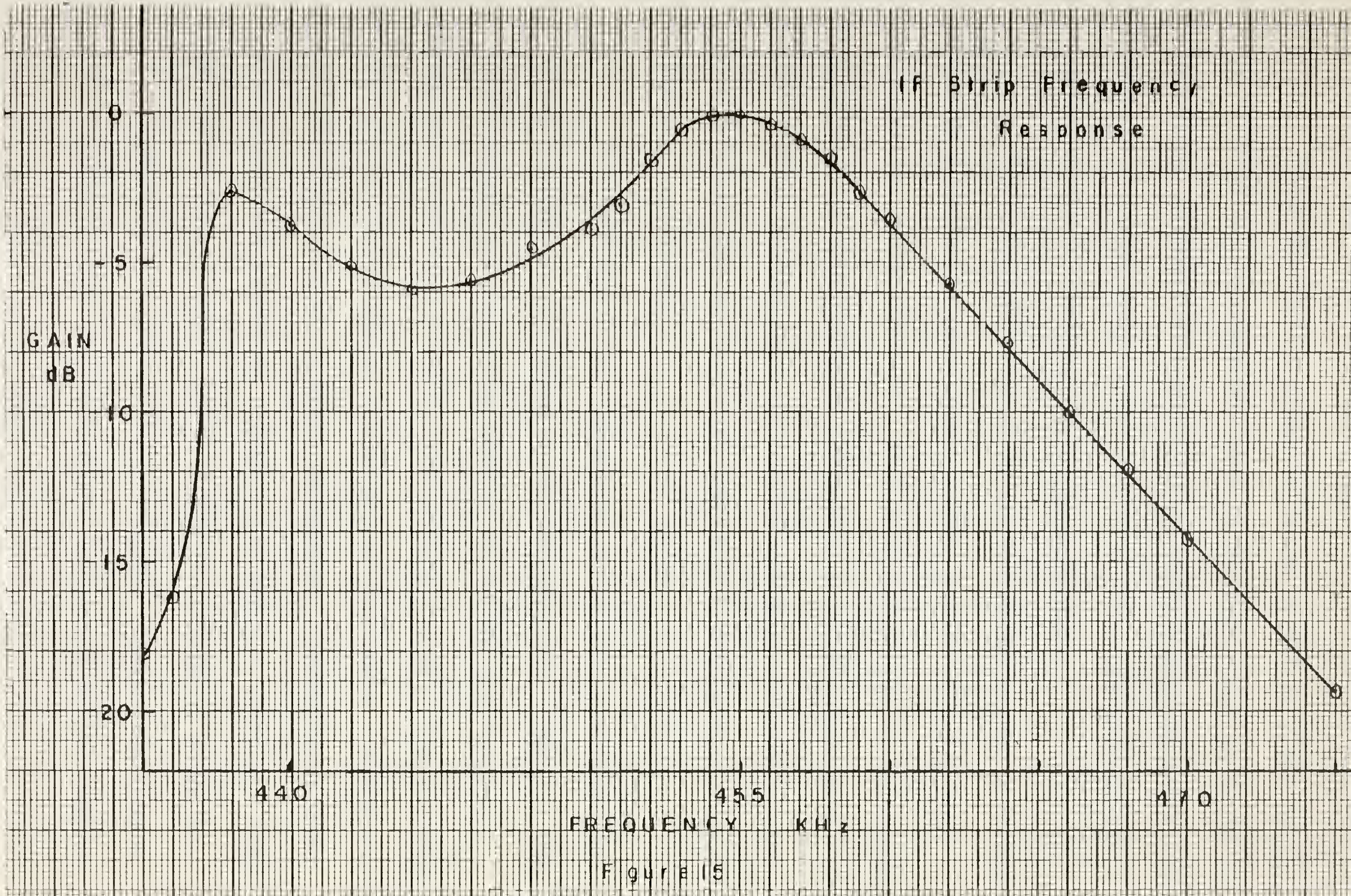



Figure 15

I. F. Strip Frequency Response

f. Audio Circuitry (Figure 16)

Since the limiter output is loaded by R65 at 1,000 ohms, coupling directly into the sensitive input of the ratio detector would be unsatisfactory. A buffer amplifier (Q12) is used to pass the signal from the limiter to the ratio detector for detection. The dual gate MOSFET is connected in triode, common source configuration.

Audio demodulation of the 455 KHz F.M. signal is accomplished using a miniature ratio detector (T5), manufactured by Bell Industries, J.W. Miller Division. This double-tuned device demodulates audio satisfactorily with observed 3 dB points at 120 Hz and 4 KHz. The demodulated audio is capacitatively coupled into a low pass filter combination of R73 and C52 having a 3 dB roll off at 6 KHz. Filter output drives the audio amplifier.

A single monolithic integrated circuit (U9) amplifies the demodulated intelligence generated in the ratio detector. The LM 380 audio amplifier has gain internally fixed at 50 (34 dB) with volume adjustment accomplished by R74, a trim potientiometer. The amplifier is ground referenced and has the ability to deliver 2 watts. Audio output is capacitively coupled to the dynamic headphones. These headphones when wired in parallel exhibit an impedance of only 1.5 ohms at 3 KHz. A 10 ohm resistance is inserted prior to the headphones to match the output of the amplifier with that of the earphones. While a significant voltage drop occurs across this resistance, the gain of the OP AMP is

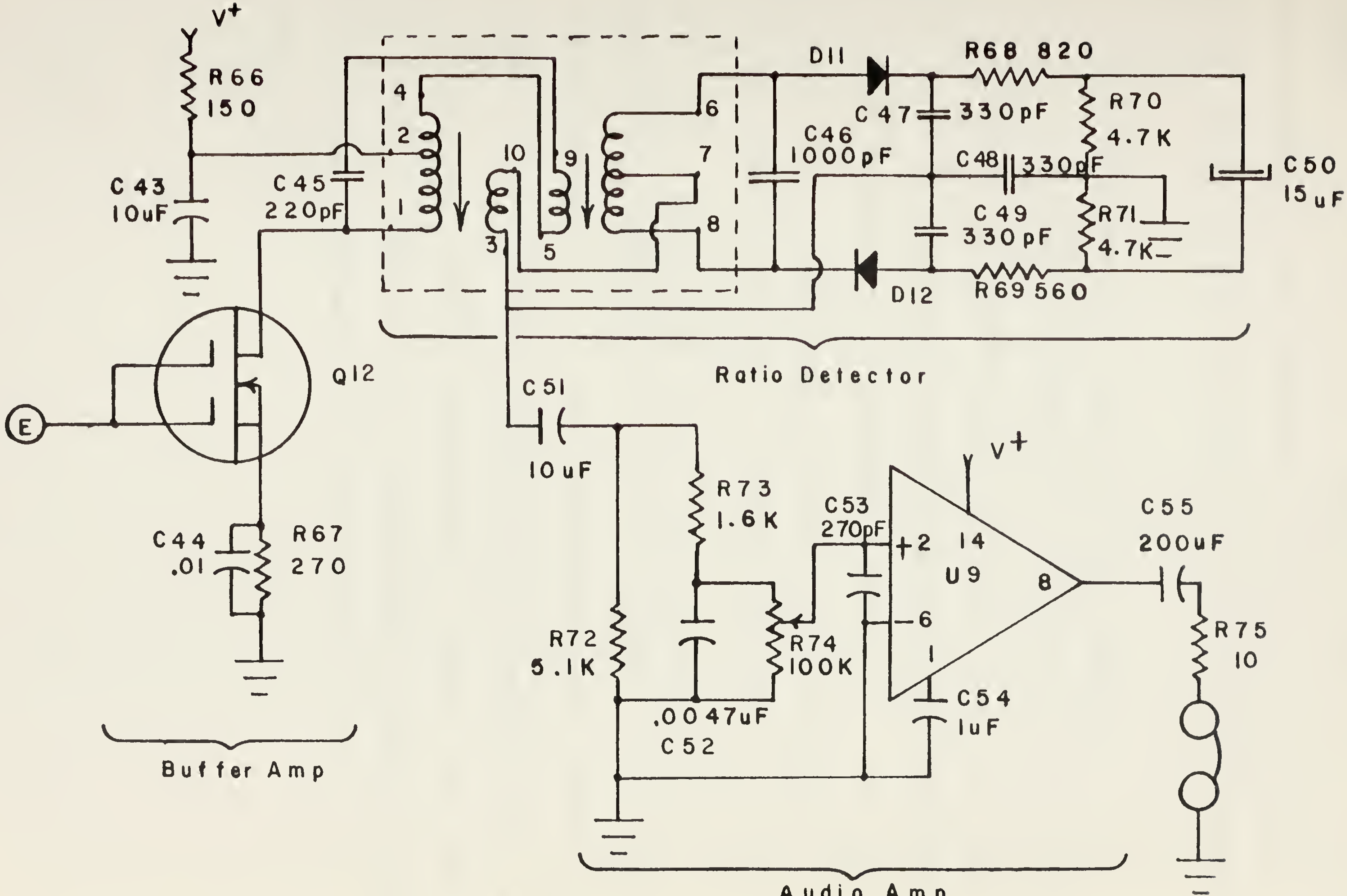



Figure 16

Schematic Diagram: Receiver Buffer Amp,
Ratio Detector and Audio Amp

more than sufficient to generate the 100 millivolts required to drive the earphones. Audio quality is good, although a somewhat more sophisticated filter arrangement at the input of the audio amplifier may enhance the quality. Buffer amplifier, ratio detector and audio amp are depicted in figure 16.

g. Automatic Gain Control Circuitry (Figure 17)

A simple arrangement for controlling gain at the receiver pre-amplifier was designed using the signal at the output of the limiter, pin 6, as a source. When the voltage at this pin reaches 280 millivolts, the onset of limiting occurs at the limiter bridge output. D13 is germanium, and will conduct at nearly the same time as the onset of limiting. This diode allows for a positive threshold below which there is no A.G.C. voltage applied to the pre-amp. R77 and C57 form a low-pass filter combination with 3 dB point at 100 Hz. Output at point "F" is slightly positive at .2 volts under grounded input conditions. Voltages of 2.5 volts negative is within the circuit capability while 1 volt negative is sufficient to attenuate the R.F. signal by more than 20 dB at the pre-amplifier. A.G.C. amplifier gain is adjustable within limits using the potentiometer, R79. Gain versus A.G.C. voltage is plotted in figure 18.

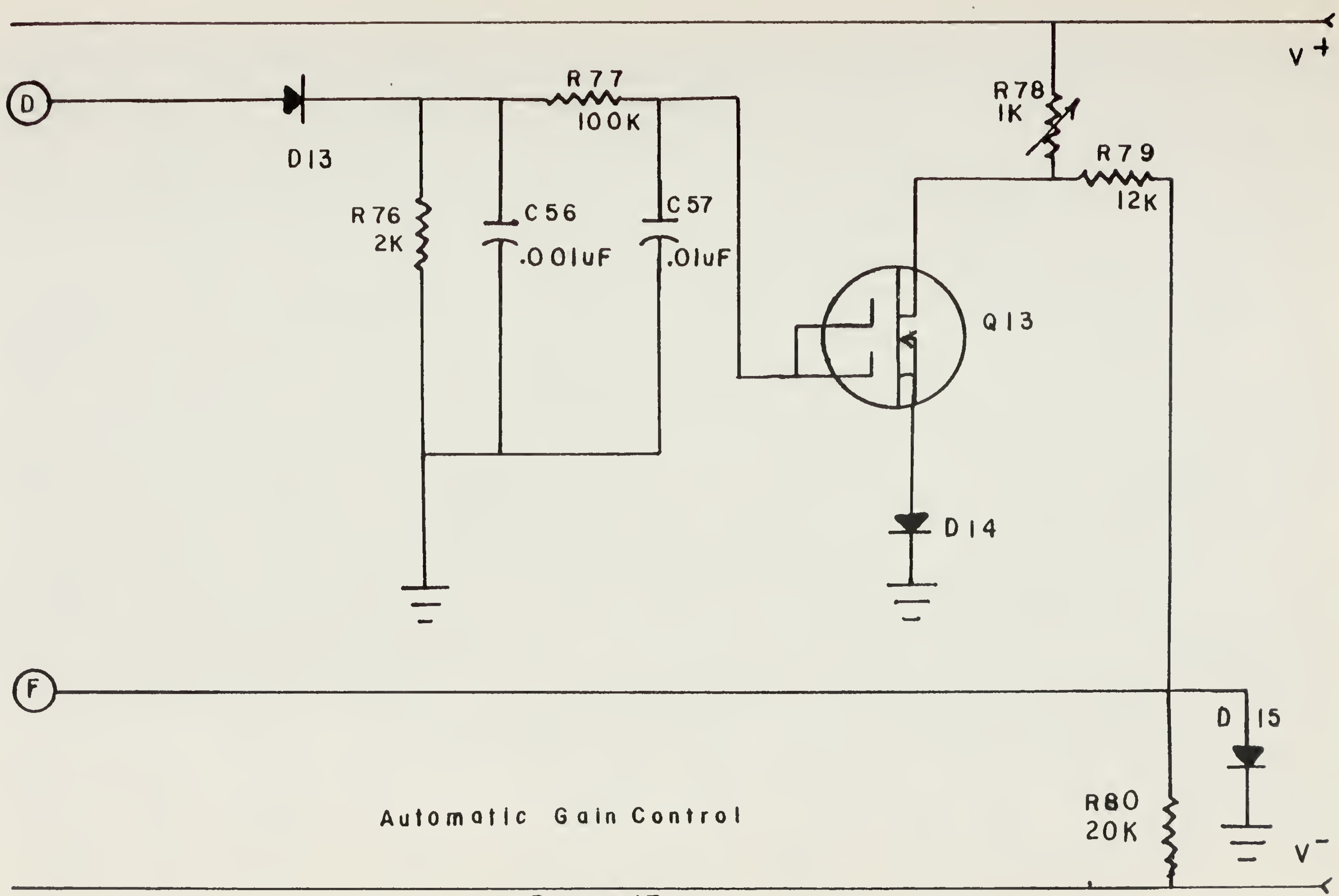



Figure 17

Schematic Diagram: Receiver Automatic Gain Control

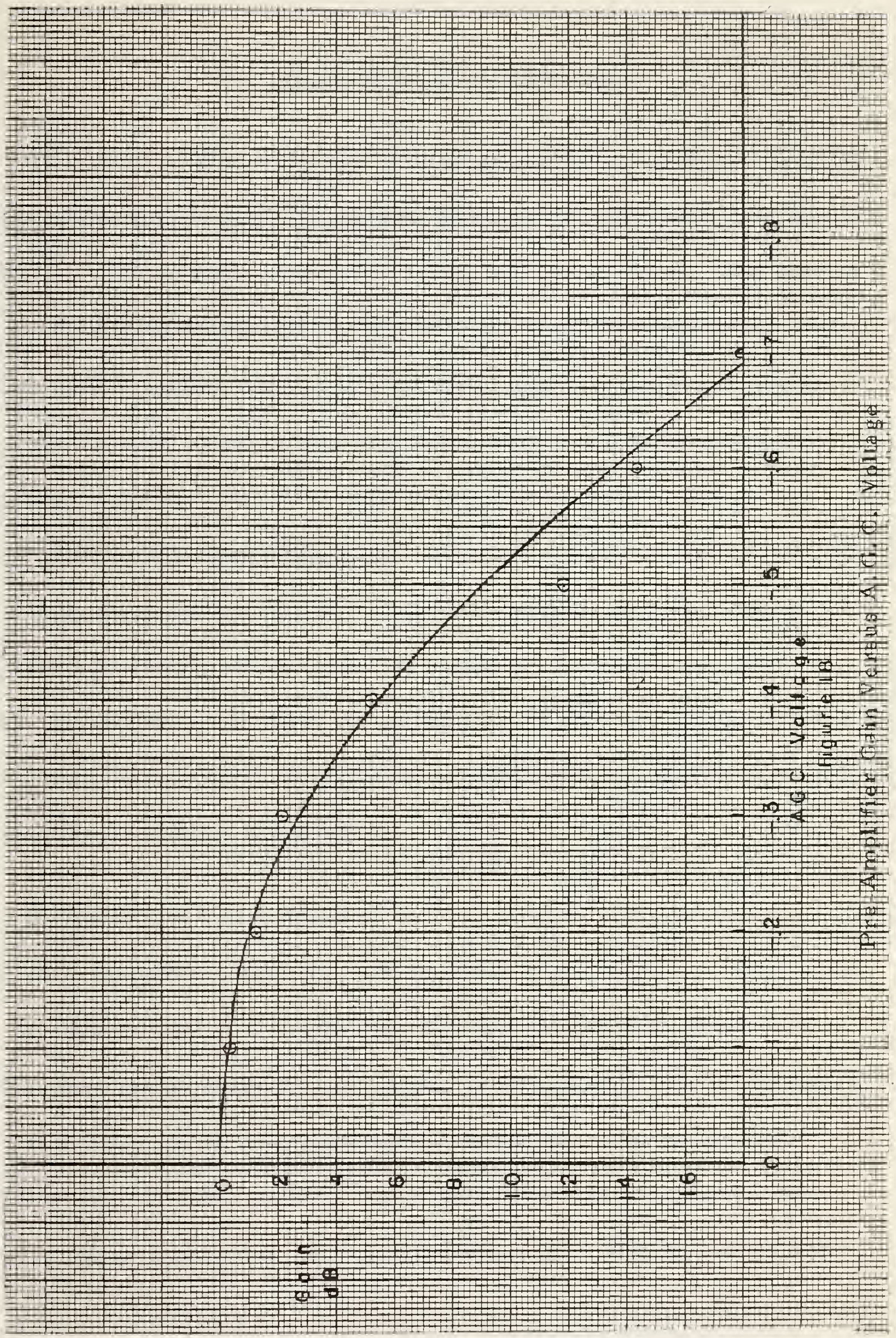



figure 18

Pre Amplifier Gain Versus A. G. C. Voltage

# IV. UNDERWATER ACOUSTICS

The intent of this section is to supplement the underwater acoustics research earlier done by Lieutenant Steere in his construction of DUCS I [Ref. 2]. Two specific objectives are to be accomplished. The first is to set forth a mathematical model which matches the impedance of the piezoelectric ceramic transducer element such that maximum power transfer into the water is effected. The results are to be physically realized in DUCS II and then empirically substantiated. The second requirement is to determine graphically the radiation pattern presented by DUCS II around both the horizontal and vertical axes in order to evaluate effectively the system. Attempts to communicate with a receiver located 10 dB down in the pattern would be an obvious waste of the time. These two considerations are preceded by a physical description of the transducer element.

## A. TRANSDUCER ELEMENT

The transducer elements used in DUCS II are similar to those used by Lt. Steere and were supplied by Mr. Ted C. Madison of Channel Industries, Santa Barbara, California. A dimensional representation as installed in DUCS II is presented in figure 19. The transducer element is a ceramic crystal ring of lead zirconate composition with bonded silver conducting surfaces inside and out. Height, inside and

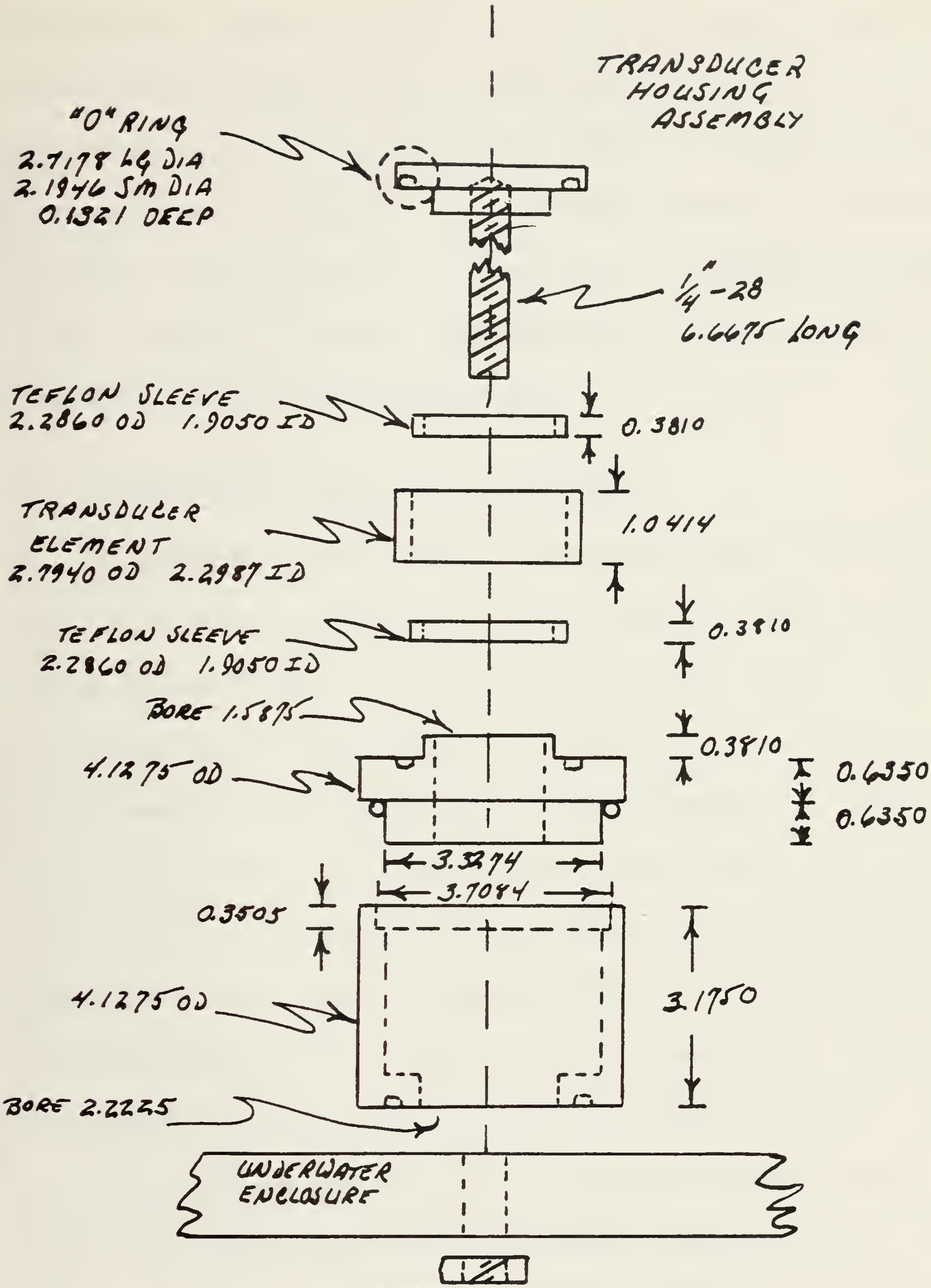


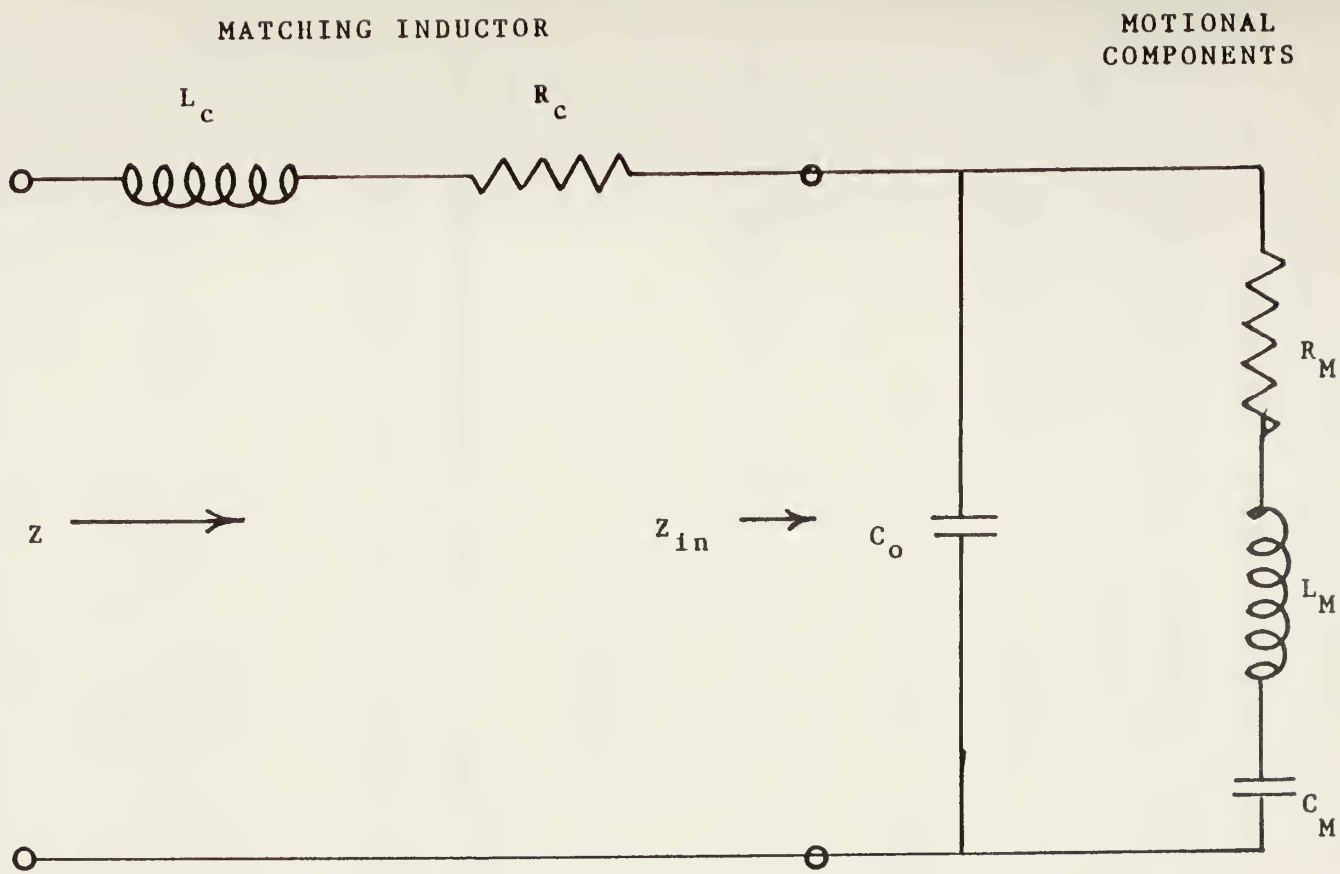

Figure 19

Transducer Mounting Assembly

outside diameter dimensions are 1.041, 2.299 and 2.794 centimeters respectively. The inner electrode is silver soldered to two fine, opposing multifilament leads which couple into the transmitter and receiver circuitry as shown in figures 8 and 11. The outer electrode is ground referenced and the element and the mounting assembly is coated with neoprene. Isolation of the inner electrode is accomplished with teflon buffer rings and "O" rings. The transducer element, as driven, expands and contracts in the radial mode.

## B. TRANSDUCER IMPEDANCE MATCHING

A mathematical model for the thin walled ceramic ring transducer element is shown in figure 20. $C_o$ is the capacitance which is formed by the concentric bonded inner and outer silvered surfaces of the ceramic element. This value is easily measured on an impedance bridge when driven at a frequency well off of resonance, since $R_m$ may be considered open. At 1 KHz, the capacitance was determined to be $3.27 \times 10^{-9}$F. $R_m$, $L_m$ and $C_m$ represent the electrical equivalent components of the crystal's complex mechanical impedance as discussed by Leon Camp [Ref. 7]. $L_c$ and $R_c$ are to be the values as determined by the impedance matching coil and its associated resistance. The transducer element was mounted in the transducer test assembly, immersed in water and driven by a variable frequency source. Using a Drametz Engineering Laboratories model 100B complex impedance




TRANSDUCER EQUIVALENT CIRCUIT

Figure 20

meter and an X-6 plotter, a graph of resistance versus reactance as a function of frequency was obtained. At a loaded resonance of 40 KHz, the resistive value was 410 ohms while the capacitive reactance value was 920 ohms. Having determined the value for $C_o$, and the values for both the resistive and reactive components of $Z_{in}$, several alternative methods may be used for fixing the value of $L_c$. Considering first the series impedance of the motional components:

$$Z_m = R_m + j\left(\omega L_m - \frac{1}{\omega C_m}\right)$$

Combining parallel impedances gives:

$$Z_{in} = \frac{Z_m}{1 + j\omega C_o Z_m}$$

Substituting the expanded value for $Z_m$, simplifying and rationalizing, the general expression for $Z_{in}$ at an arbitrary frequency near resonance, , is:

$$Z_{in} = \frac{R_m + j\left(\omega L_m - \frac{1}{\omega C_m}\right)\left[\left(1 - \omega L_m C_o + \frac{C_o}{C_m}\right) - j C_o R_m\right]}{\left(1 - \omega^2 L_m C_o + \frac{C_o}{C_m}\right)^2 + \omega^2 C_o^2 R_m^2}$$

At mechanical resonance $X_{mc}$ and $X_{ml}$ are equal and opposite,

$\omega_{res} L_m - \frac{1}{\omega_{res} C_m} = 0$. Rearranging the last expression for $Z_{in}$

to produce $(\omega L_m - \dfrac{1}{\omega C_m})$ and canceling these terms due to resonance yields the following much simpler expression for $Z_{in}$ at resonance:

$$Z_{in} = \frac{R_m(1 - j\omega_r C_o R_m)}{1 + \omega_r^2 C_o^2 R_m^2} = \frac{R_m}{1 + \omega_r^2 C_o^2 R_m^2} - \frac{j\omega_r C_o R_m^2}{1 + \omega_r^2 C_o^2 R_m^2}$$

Here the resistive and capacitive reactance terms previously determined for $Z_{in}$ are 410 and 920 ohms respectively. To bring the power factor closer to unity, an inductive reactance equal to the capacitive reactance is required. Thus

$$j\omega_r L_c + \left(\frac{-j\omega_r C_o R_m^2}{1 + \omega_r^2 C_o^2 R_m^2}\right) = 0$$

and

$$L_c = \frac{C_o R_m^2}{1 + \omega_r^2 C_o^2 R_m^2}$$

The impedance matching inductor, $L_c$, is finally evaluated after determining the value for $R_m$. This is done using the $Z_{in}$ equation as a source for two possible quadratics:

$$\frac{R_m}{1 + \omega_r^2 C_o^2 R_m^2} = 410 \quad \text{and} \quad \frac{\omega_r C_o R_m^2}{1 + \omega_r^2 C_o^2 R_m^2} = 920$$

Solving the quadratic equation using the quadratic formula yields two roots $R_{m1} = 3.085 \times 10^3$ ohms (sum) and $R_{m2} = 472.8$ ohms (difference). Solving for $L_c$ with these values yields $L_{c1} = 3.41$ mH (valid) and .634 mH (discarded). A similar procedure was carried out using the reactive component of $Z_{in}$ to define $R_m$. $R_m$ was $2.82 \times 10^3$ and $L_c$ then followed at 4.03 mH. Experimentation to minimize reactive power inefficiency found 4.4 mH to produce best matching. The result of this effort makes the system resonant and nearly resistive at 515 + j15 ohms as shown graphically in figure 21. Figure 22 is the graph of the system driven in air vice water to show the significant effect of mechanical loading by the water.

## C. FREE FIELD RADIATION PATTERN

In order to effectively use DUCS II, a knowledge of the radiation pattern is required. According to Steere [Ref. 2] the ceramic ring element working in the radial mode produces omni-directional coverage in one plane. This is essentially true as will be shown, however, vertical displacement from the horizontal plane has a great effect on the ability to receive signals.

### 1. Procedures

A U.S.R.D. type F27 calibrated transducer was suspended at a depth of one meter in a large anechoic water tank. The DUCS II transducer element with matching inductor was mounted on the test

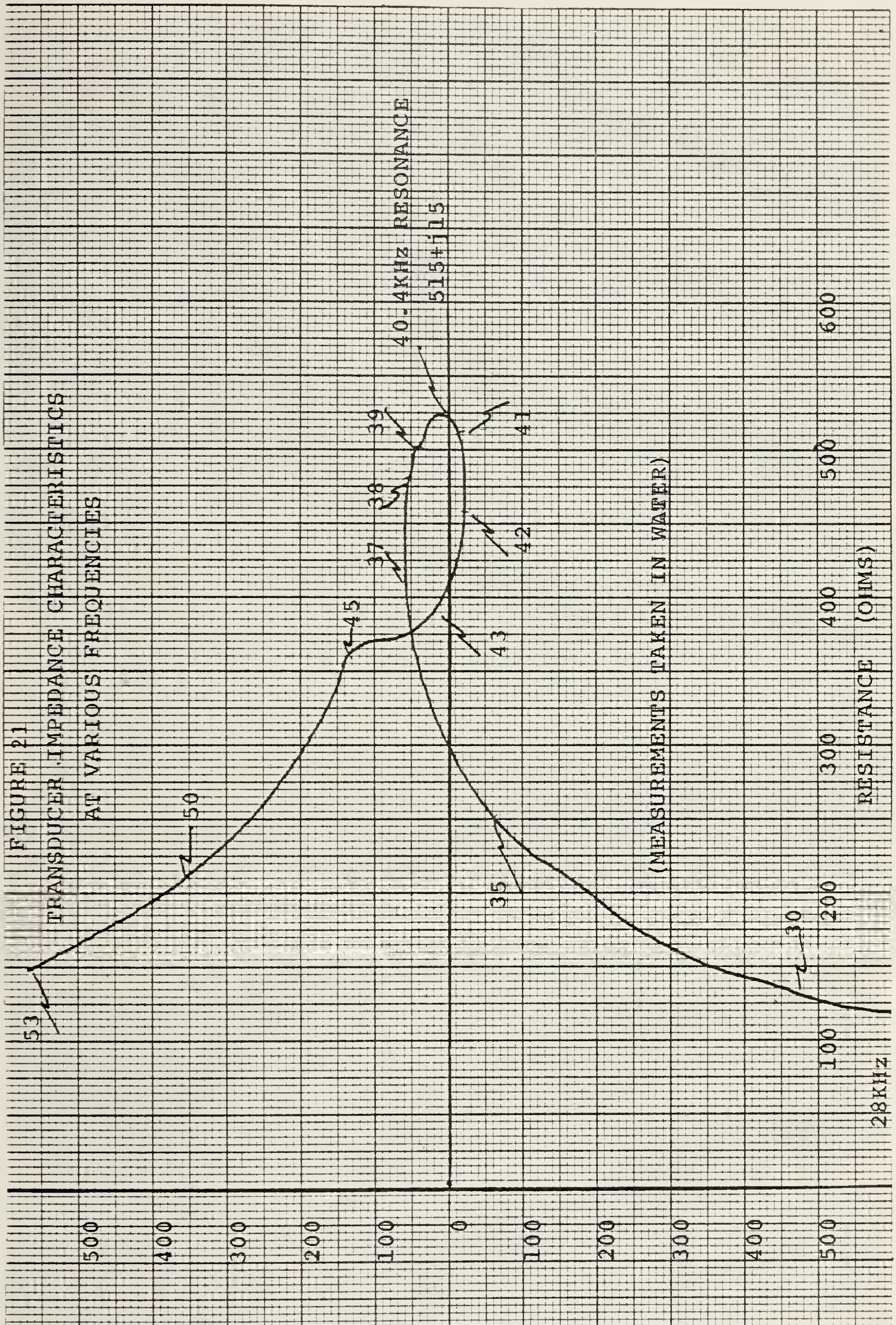


FIGURE 21
TRANSDUCER IMPEDANCE CHARACTERISTICS
AT VARIOUS FREQUENCIES
(MEASUREMENTS TAKEN IN WATER)
40.4KHz RESONANCE
515+j15
28KHz
30
35
37
38
39
41
42
43
45
50
53
100
200
300
400
500
600
RESISTANCE (OHMS)
500
400
300
200
100
0
100
200
300
400
500

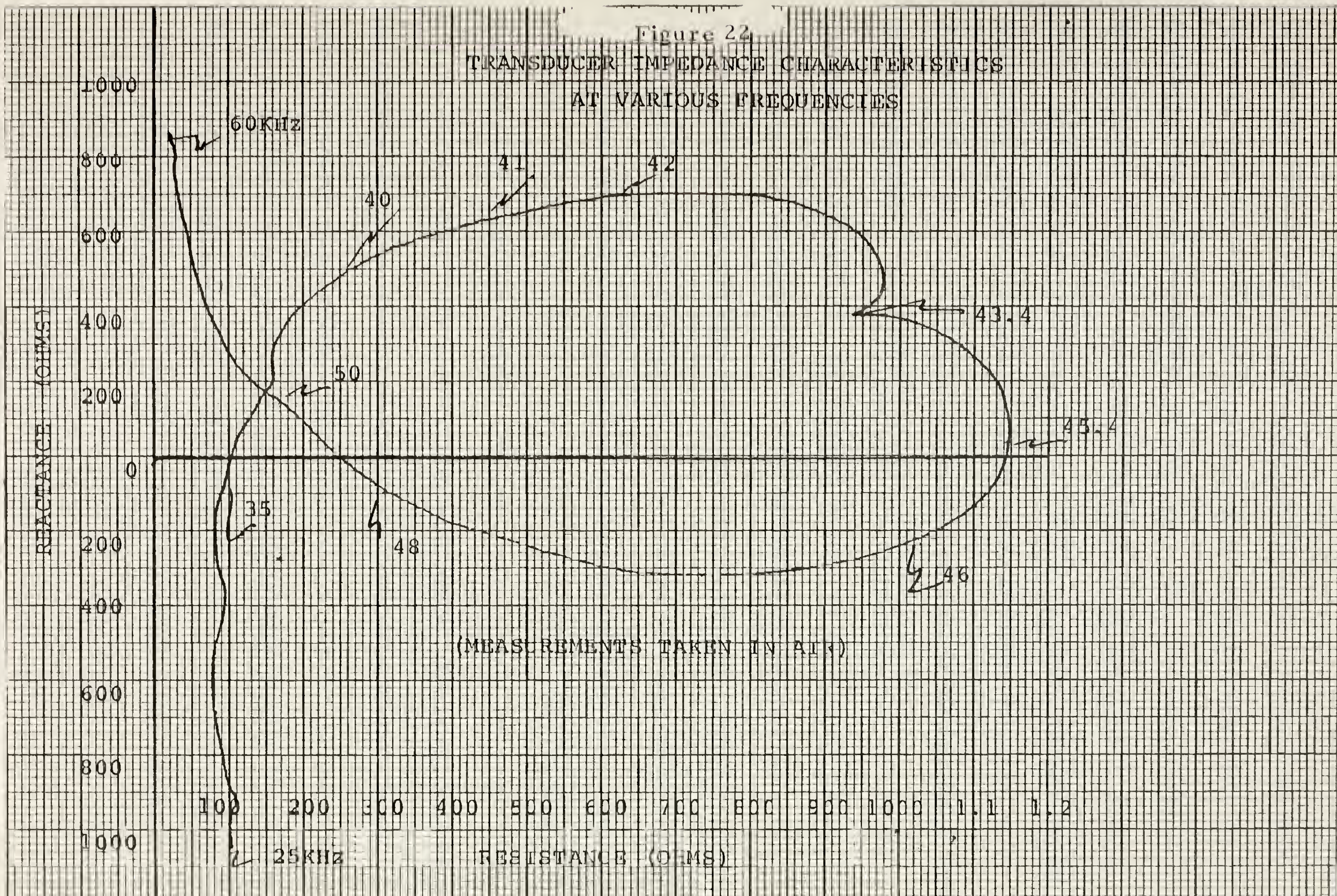


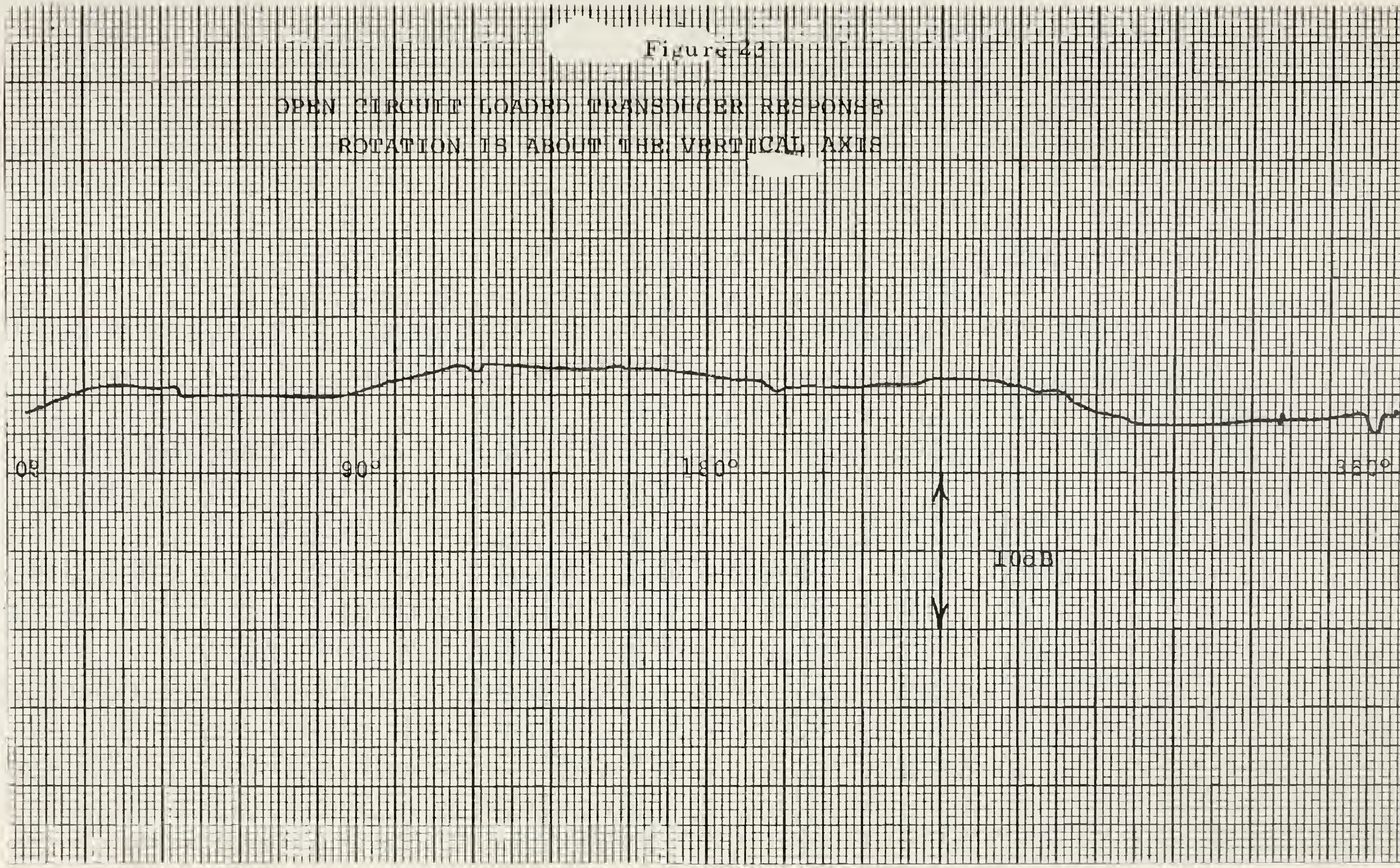

Figure 22
TRANSDUCER IMPEDANCE CHARACTERISTICS
AT VARIOUS FREQUENCIES

platform and positioned 3.5 meters from the F27 at the same depth. At the opposite end of the rod mounting the DUCS II transducer was located a synchro motor for rotation and a synchro transmitter to provide angular displacement information. The F27 was driven at 40 KHz while observing the amplified voltages of the DUCS II transducer in the hydrophone mode. A plot of the received voltages as the transducer was turned through 360 degrees was made using an X-Y plotter. Figure 23 is the graph of gain in dB versus displacement about the vertical axis. The maximum difference between any two points is something less than 5 dB; the general pattern is essentially omnidirectional. A second graph, figure 24, was plotted after mounting the transducer test platform perpendicular to the rod. Rotation about the transducer horizontal axis as plotted shows that pattern is significantly degraded off of a plane passing through the diameter of the ring element. To the right of the zero degree position is represented rotation away from the inductor housing. Twenty degrees either side of zero is less than 3 dB from zero displacement. Rotating away from the inductor housing, there is a major dip of 20 dB and dips thereafter of greater magnitudes. There is a null at the end-on position where the attenuation is down 30 dB. Performance turning away from the housing is generally better than rotating toward it. Assuming a sound velocity of $1.5 \times 10^5$ cm/sec, the wavelength of the radiated

Figure 23

OPEN CIRCUIT LOADED TRANSDUCER RESPONSE

ROTATION IS ABOUT THE VERTICAL AXIS



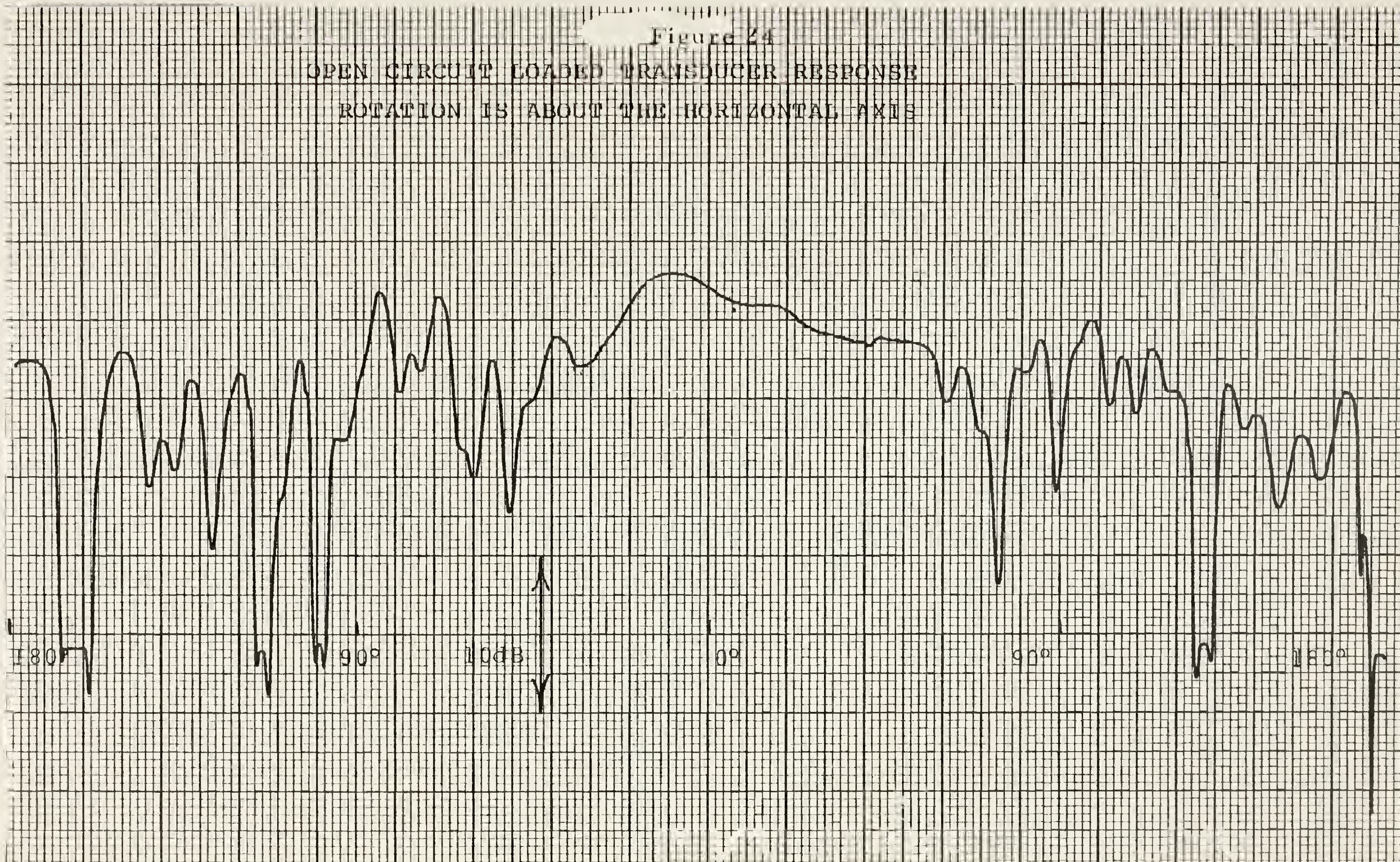



Figure 24

OPEN CIRCUIT LOADED TRANSDUCER RESPONSE

ROTATION IS ABOUT THE HORIZONTAL AXIS

energy at the resonant frequency is about 3.5 cm. The transducer housing (figure 19) is 4.1 cm. in height. The existence of destructive interference being more significant at depressed angles is to be expected. The radiation pattern could be much improved with reconfiguration of the mounting assembly.

# V. SUPPORTING EQUIPMENT

Any attempts to design an electronics package for the purpose of diver communication may be invalidated by supporting equipment which is uncomfortable or which is incapable of effectively coupling input and output from the electronics package to the diver. An effort has been made to design or obtain peripheral equipment for DUCS II which enhances effective use of the electronic and ultrasonic components and meet the requirement of low cost.

## A. MICROPHONE

The microphone in DUCS II was provided by Mr. Larry Dewberry of the Naval Coastal Systems Laboratory, Panama City, Florida, where it was manufactured. Although an actual frequency response test was not made, a generalized curve provided by N. C. S. L. is presented in Figure 25. This microphone is a series-connected, bimorphic ring type, using a bimorph disk as the primary transducer. It exhibits a capacitance of .015 uF. The ceramic composition is lead zirconate titanate. The transducer is ring mounted and air backed, thus enjoying pressure equalization with no distortion generated at the microphone as demand pressure from scuba increases with depth. The transducer of 2.24 cm is set off from the base on a ring of 1.91 cm inside diameter. Figure 26 is a photograph showing both the top

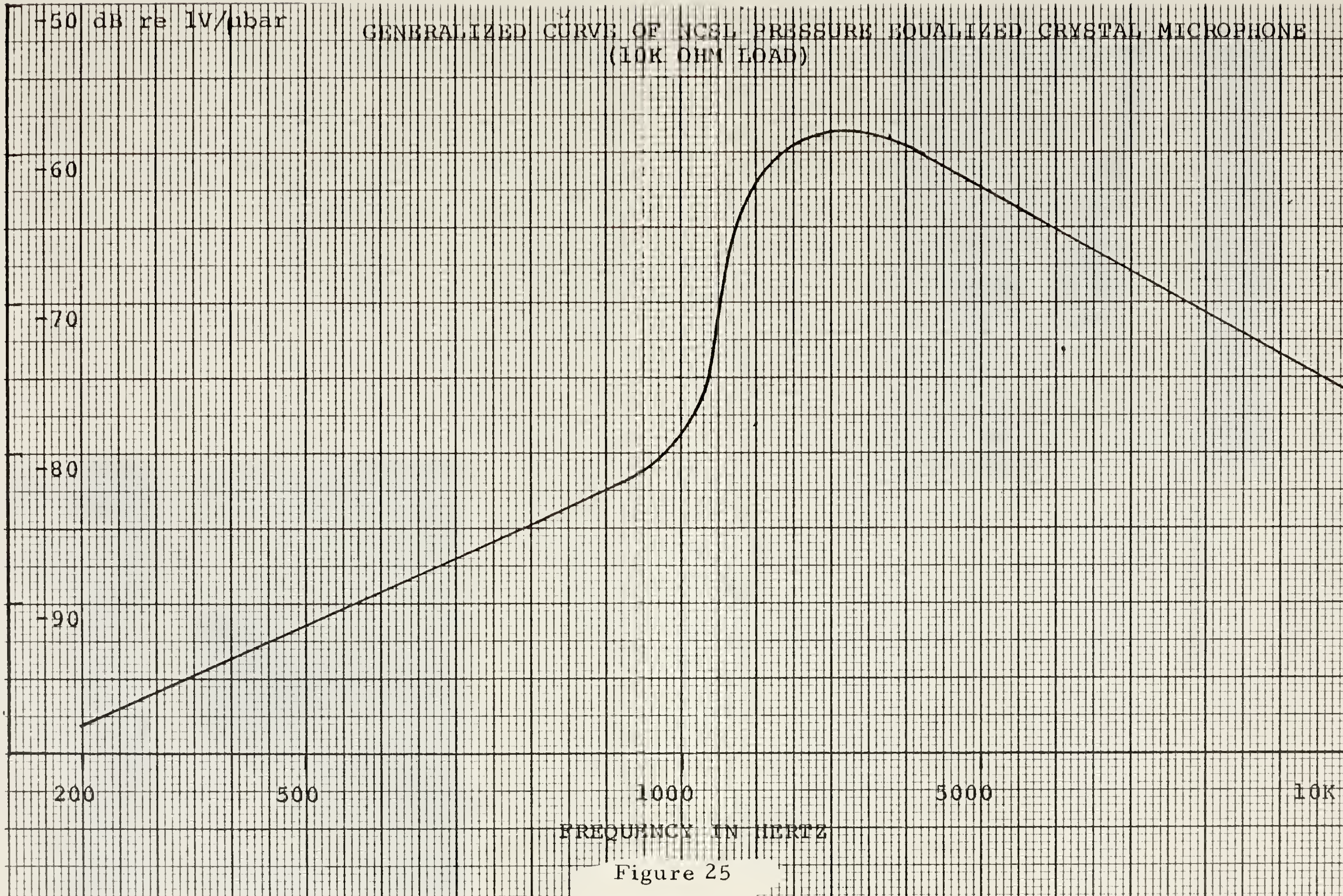
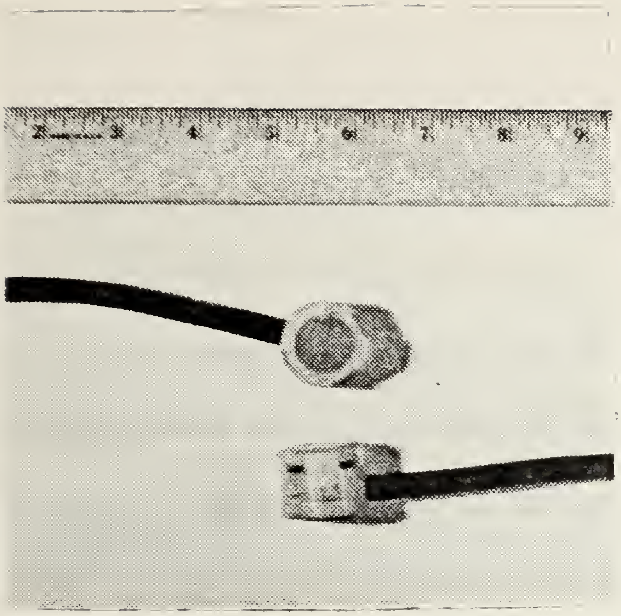

GENERALIZED CURVE OF NCSL PRESSURE EQUALIZED CRYSTAL MICROPHONE
(10K OHM LOAD)
-50 dB re 1V/μbar
-60
-70
-80
-90
200
500
1000
5000
10K
FREQUENCY IN HERTZ

Figure 25

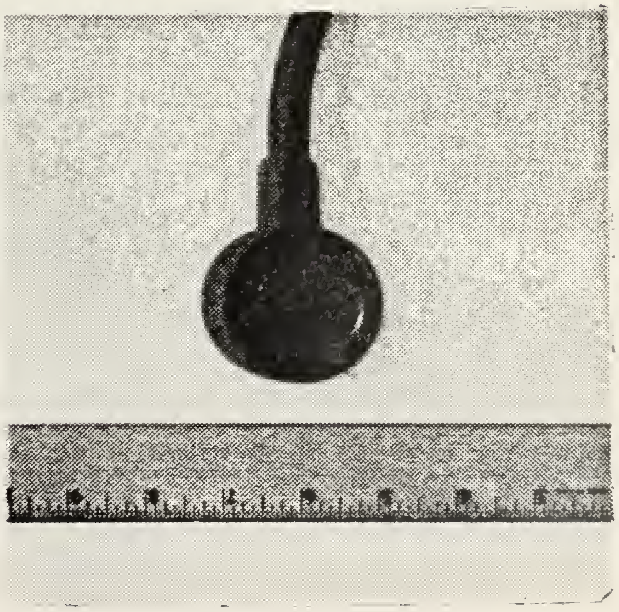


**Microphone**

Figure 26

**Earphone**

Figure 27



and profile views.

## B. EARPHONES

Earphones selected for DUCS II are of the dynamic or moving coil type. They offer a significant improvement in both comfort and effective coupling to the swimmer over the Aquasonics U420 used in DUCS I. Steere utilized the bone conduction mode of coupling in placement over the mastoid and temporal areas of the head. The inability to anchor the earphones led to intermittent reception.

Initial considerations of bone conduction coupling in DUCS II were abandoned after discussions with Mr. Bev Morgan of Deep Water Development Corporation indicated that excellent results were to be obtained by placing the earphones directly over the ear. This method was tried using earphones distributed by U. S. Divers Corporation of Santa Ana, California (3904 series). The two inch diameter earphones are rubber coated and distribute the pressure of the wet suit well. They have a satisfactory frequency response. When tested underwater, 200 millivolts constant amplitude afforded good audio. Disadvantages include no pressure equalization and low impedance. Two earphones wired in parallel were measured on an impedance bridge where a resistance of 3.5 ohms and an inductance of .52 mH was obtained. Enhanced signal quality from the audio amplifier (U9) is effected using a 10 ohm (R75) resistance in series before the headphones. The output

capability can easily overcome any voltage drop across R75. The earphones are sewn into the wet suit hood using one eighth inch neoprene material.

## C. FULL FACE MASK

A U. S. Divers full face mask, part number 5205, was selected for use in DUCS II and is pictured in figure 28. The arrow shows the Electro-Oceanics right angle connector which allows the diver to disconnect himself from the system with minimum inconvenience. The mask proved to be comfortable as tested. Jaw movement associated with articulation is reasonably free. Some difficulty was encountered in clearing of ears with descent. This unit represents a great improvement over the unit tested by Lt. Steere in both comfort and utility with respect to microphone placement and the ability to disconnect without removing the mask.

## D. UNDERWATER ENCLOSURE

An underwater enclosure to house the electronics of DUCS II was designed and is shown in figures 29, 30 and 31. Mr. Bob Moeller machined all parts and was instrumental in the design. The all aluminum unit was painted with zinc chromate primer and epoxy paint for corrosion protection. Two holes were honed and tapped for a double and single E.O. connector. A third was drilled for connection access to impedance matching inductor and transducer. All sealing

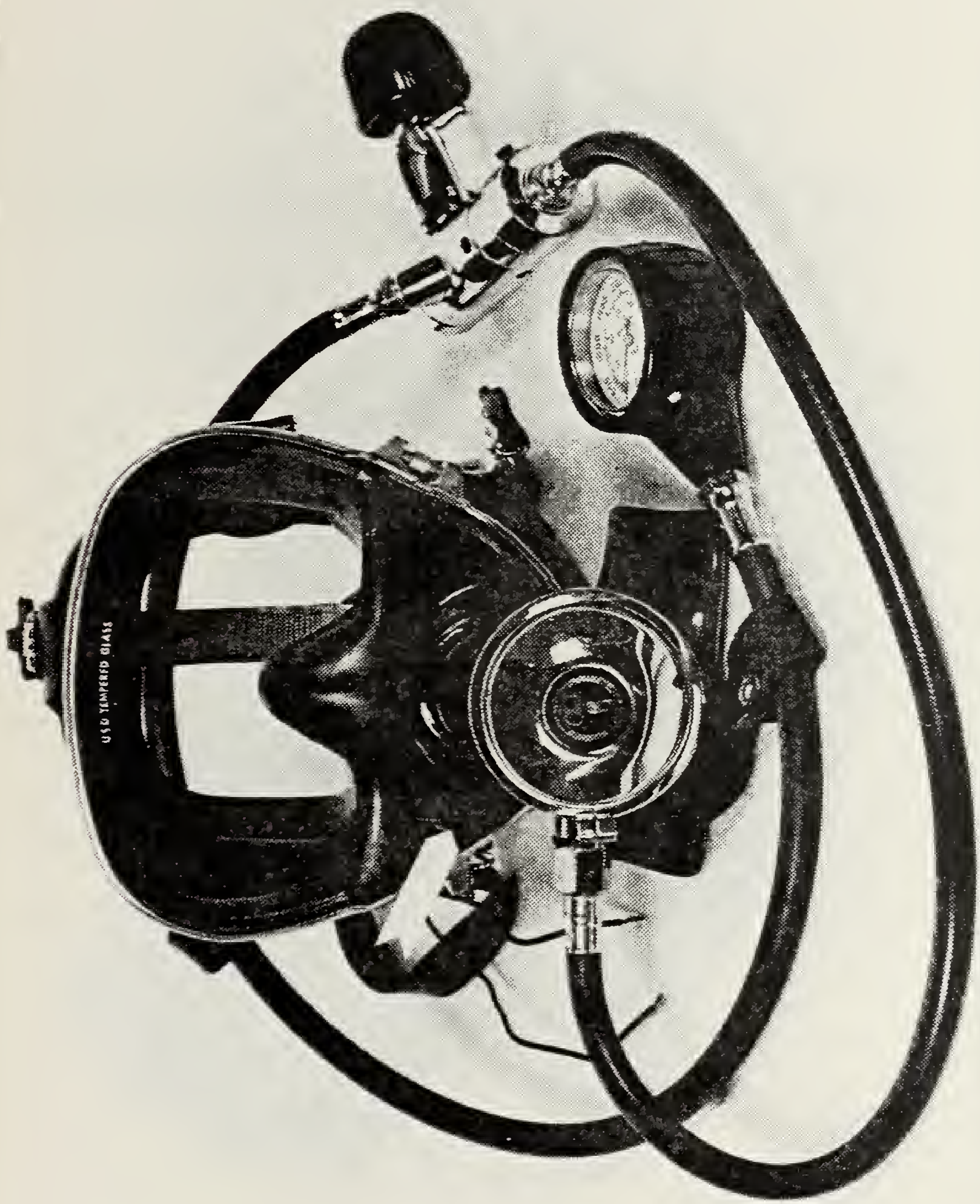



Figure 28
Full-Face Mask

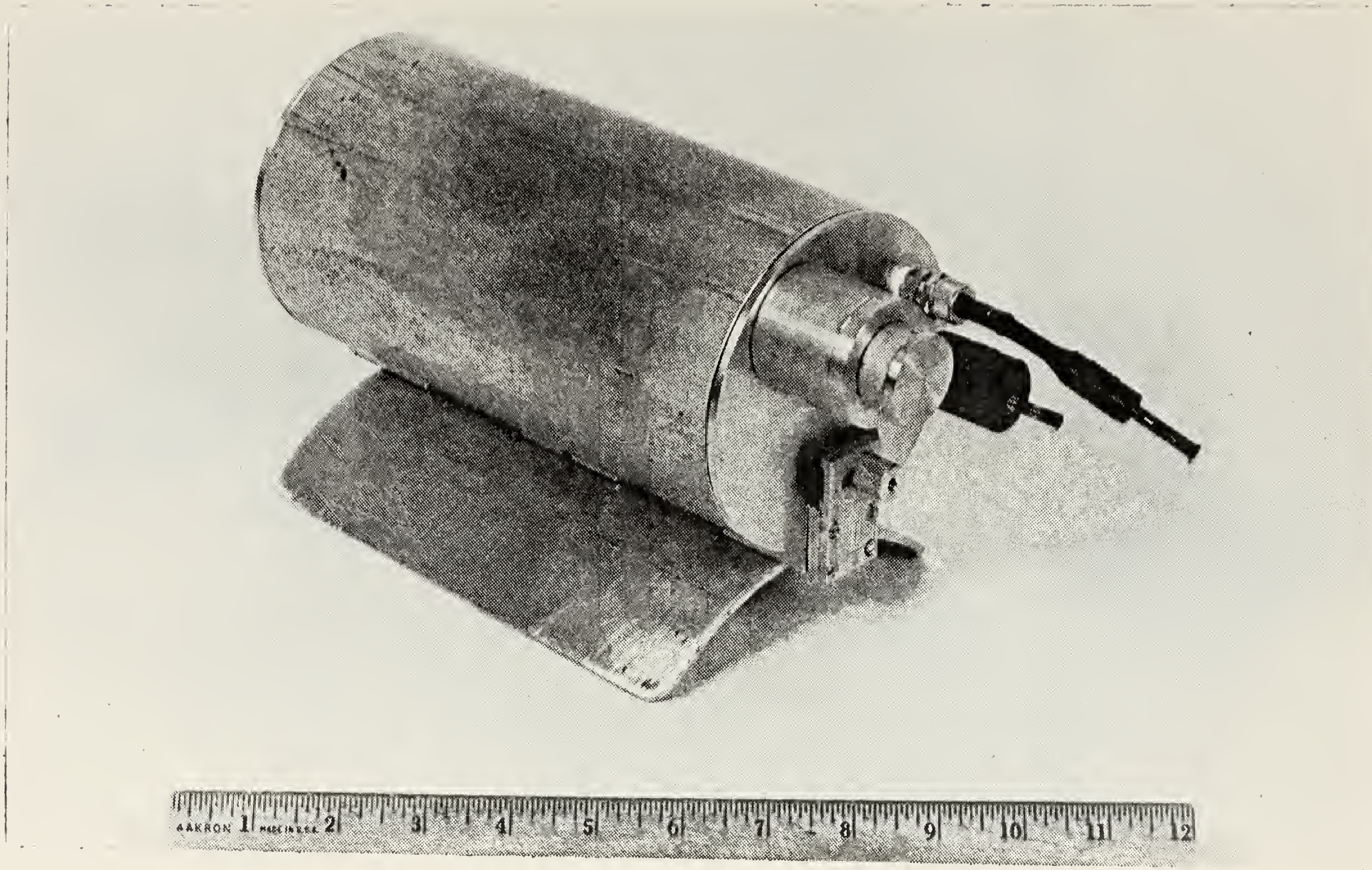



Figure 29

Underwater Enclosure, Assembled

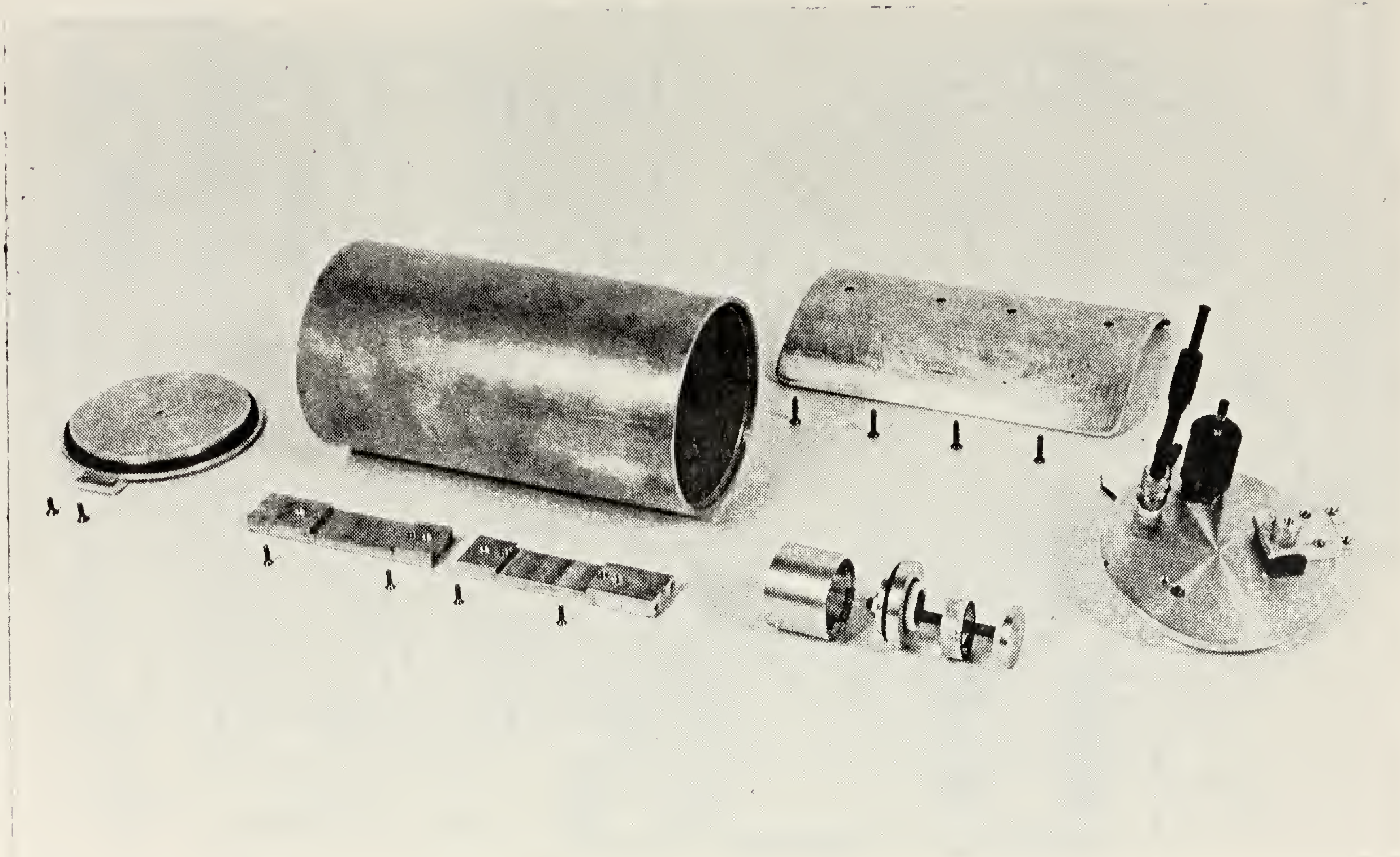


**Figure** 30
Underwater Enclosure, Exploded View

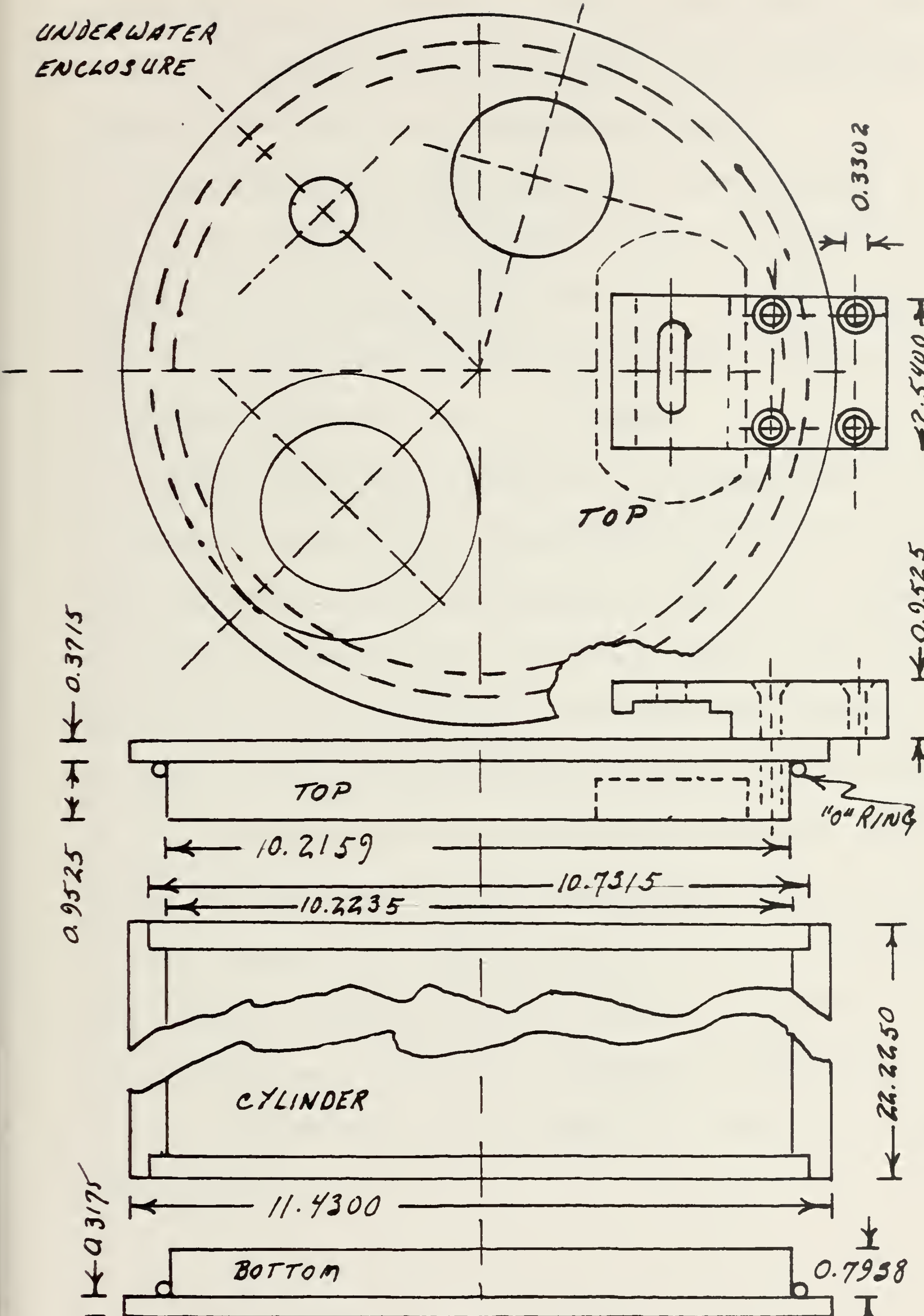



FIGURE 31

Underwater Enclosure, Dimensional View

is accomplished using "O" rings. The unit was tested to a depth of 40 meters (5 atmospheres) with no leakage. The maximum water-tight depth is expected to far exceed the test depth.

While aluminum is satisfactory for use in the prototype, corrosion is a major problem. The design is sound and could likely be used for production units, using a resilient plastic as a medium. Miniaturization was not a driving concern in prototype electronics construction; the final volume is expected to be much reduced.

Figure 32 pictures the components of the magnetic power switches. Two magnetic reed switches mounted on insulating board are fitted into a recess milled into the underside of the top. As a magnet is positioned above the switches, switching is effected passing the power supply voltages to the electronics. A separate reed switch is provided for both V+ and V-. This switching arrangement reduces the number of penetrations in the enclosure with attendant savings in additional fittings and enhancement of watertight integrity. Access to the electronics is made by inserting a screwdriver in a keyway between top and cylinder. Considerable force is required to close the unit. A simple field press consisting of a "U" shaped aluminum frame and a screw for the leverage is shown in figure 33. Operation is simple and effective.

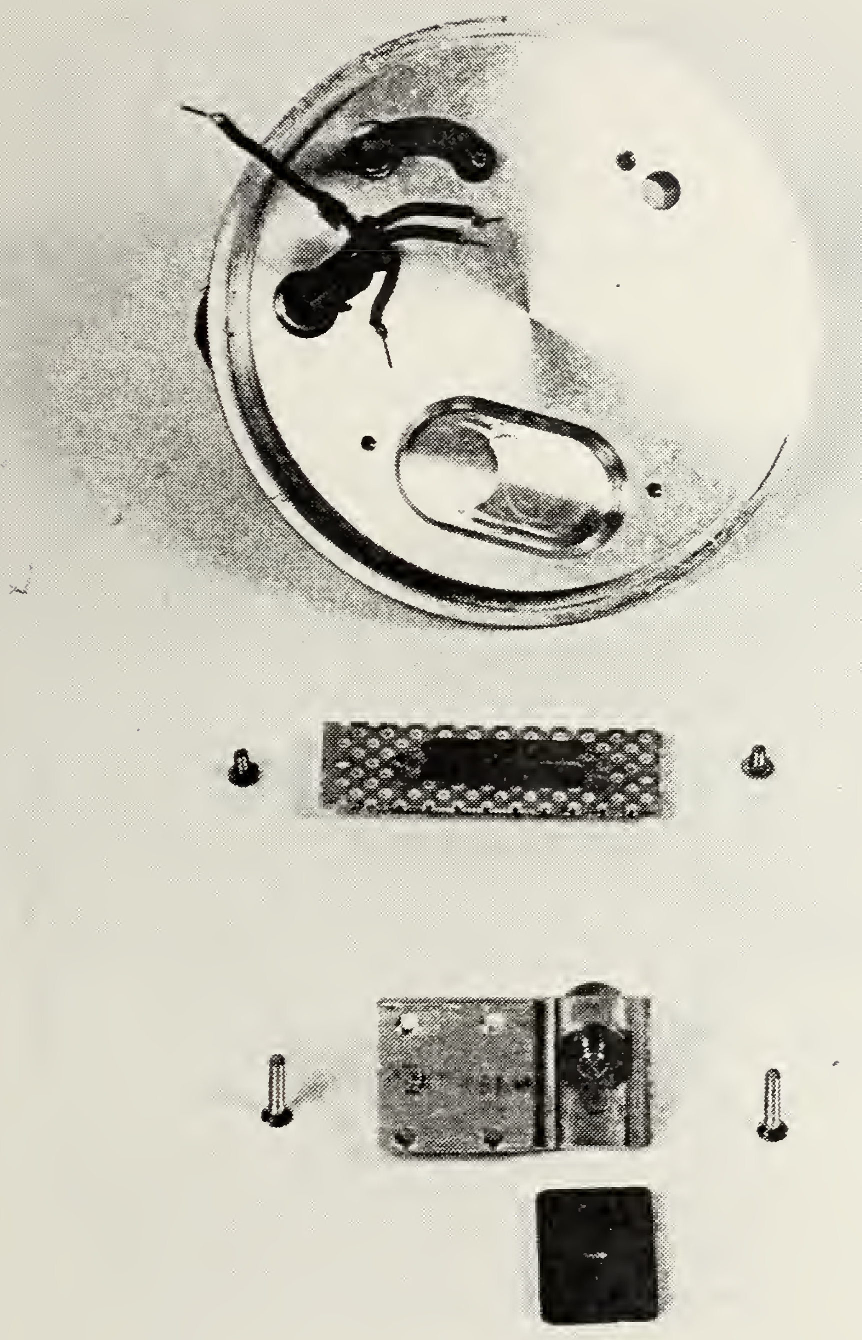
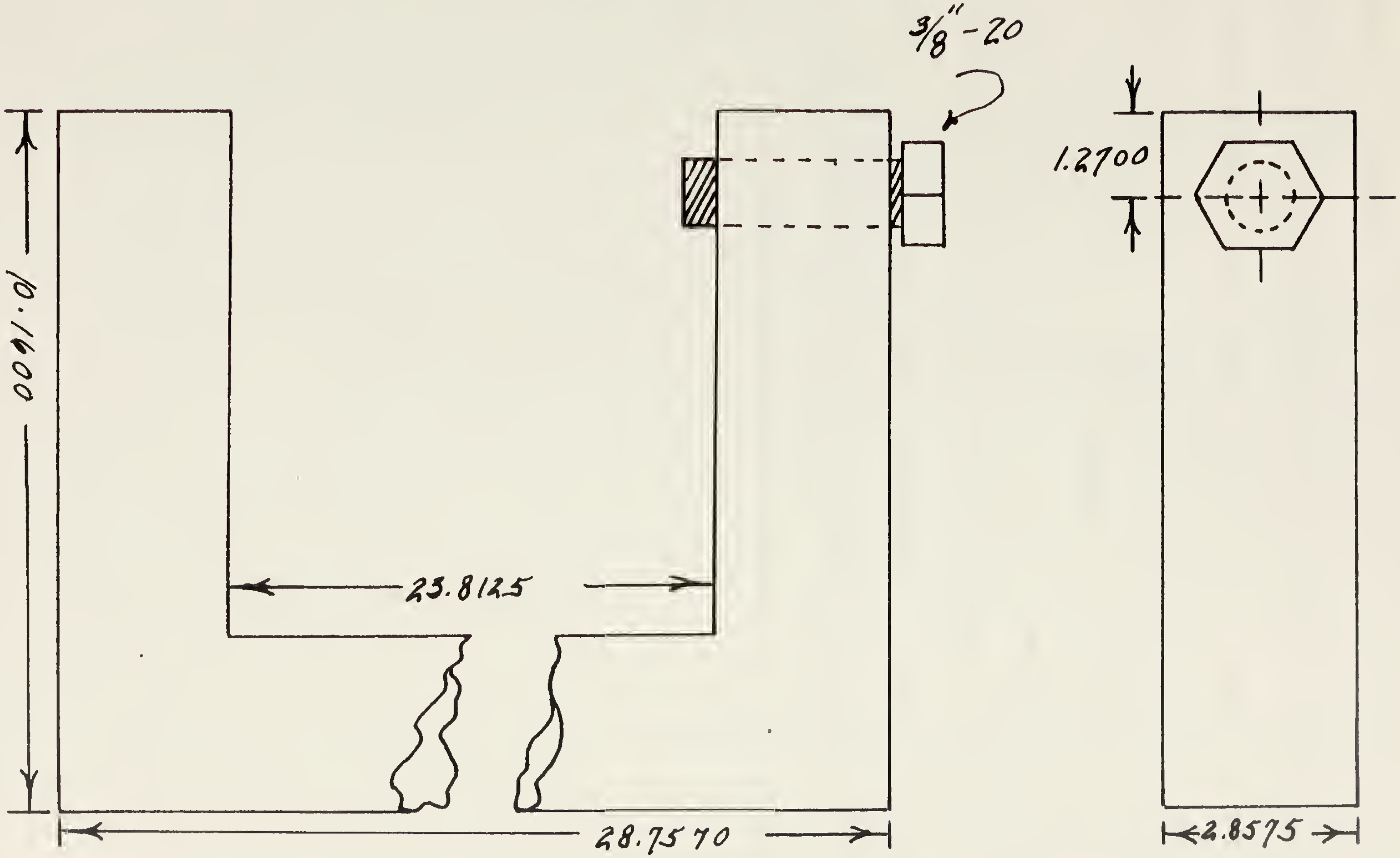


Figure 32

Magnetic Proximity Switch Assembly

FIGURE 33

FIELD PRESS UNIT



## E. TRANSDUCER TEST ASSEMBLY

During electronics design and construction, the need to determine transducer characteristics became apparent. To this end, a platform for transducer testing was fabricated and is shown in figures 34 and 35. The dimensions of the inductor housing match those of the final assembly so that radiation patterns for both the test platform and the enclosure will be similar. The component in the center of figure 35 with battery mounting brackets was included to permit addition of a broad-banded pre-amplifier within the test platform. Use of this unit in ship to diver communications is envisioned. A small pre-amplifier would be advantageous with respect to noise considerations. At depth, the unit could be lowered into the diver's beam pattern to insure good reception.

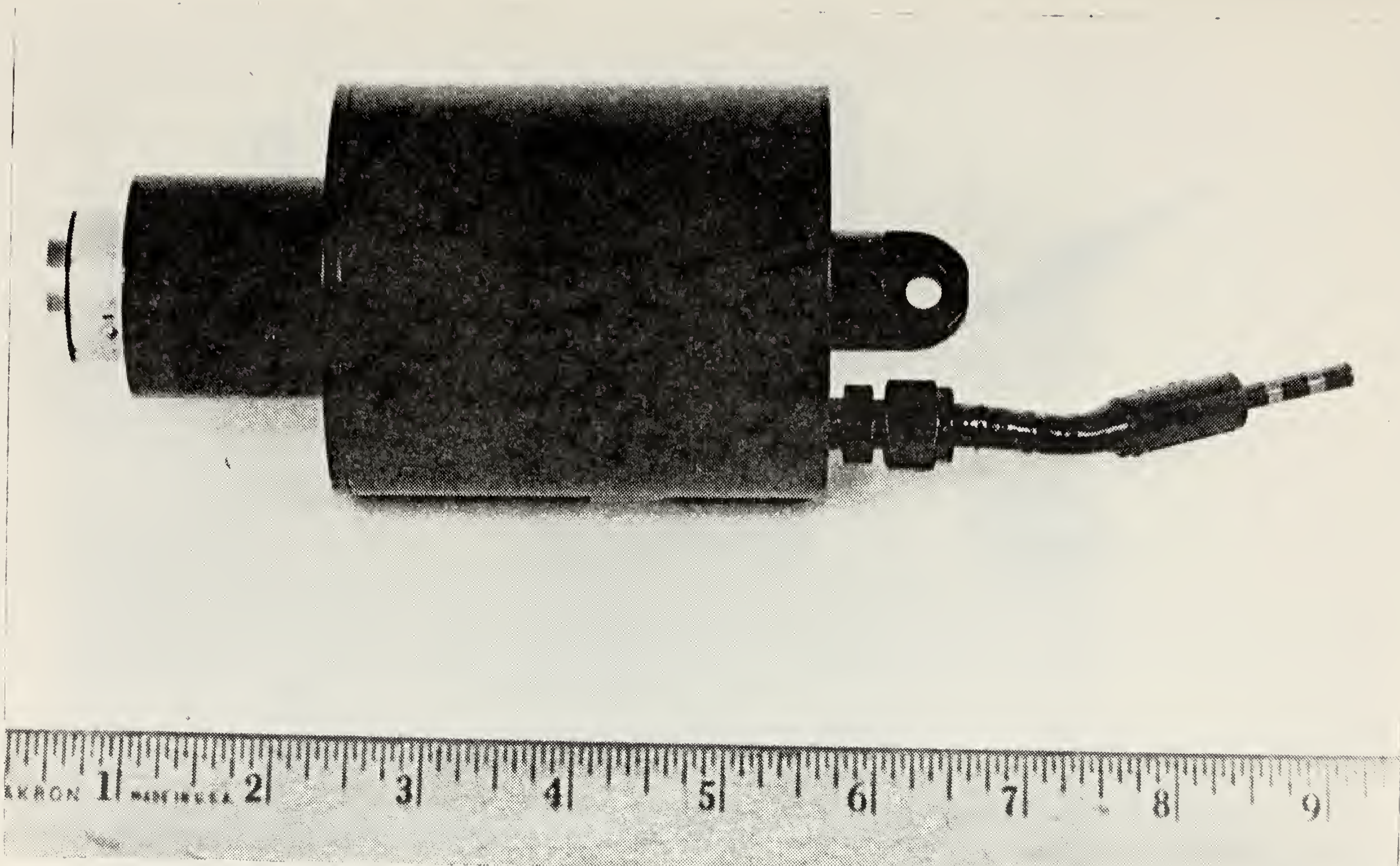



Figure 34

Transducer Test Assembly

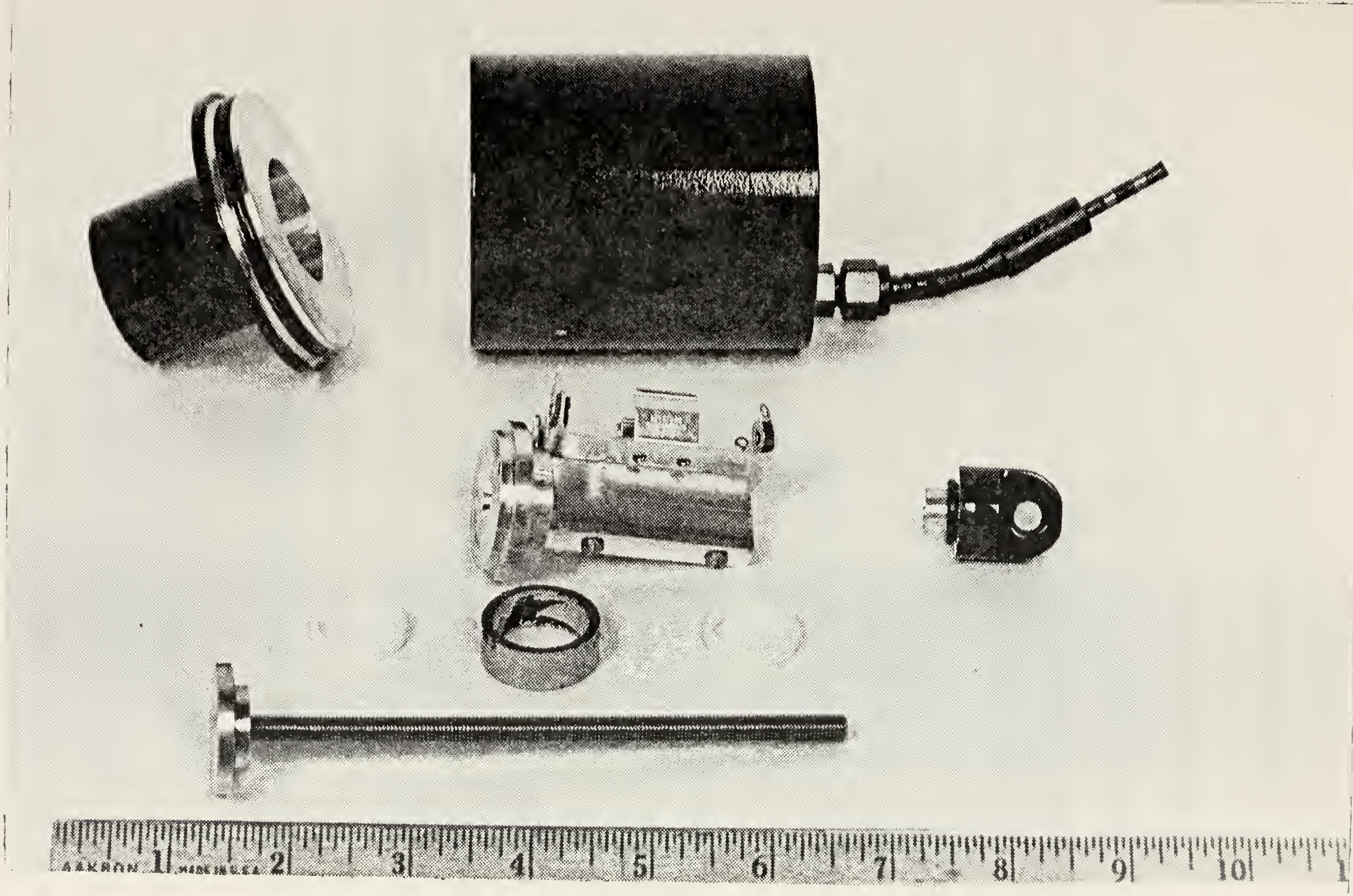


Figure 35

Transducer Test Assembly, Exploded View

# VI. CONCLUSIONS AND RECOMMENDATIONS

Several dives in the Monterey Bay area were conducted in evaluating DUCS II. However, a comprehensive and complete system evaluation is still required. All items of supporting equipment have been proven satisfactory and further effort in this specific area is not considered necessary to evaluate effectively the electronics package. Receiver sensitivity is excellent and one thousand meter range as proposed by Lt. Steere [Ref. 2] is anticipated. Results of a complete evaluation will be made available in a future report.

The first recommendation for improvement addresses the insufficient information concerning the voice transfer function of the mask and microphone. A pre-emphasis/de-emphasis network could be designed to enhance those frequencies which are attenuated in the mask. High quality, broad-banded, portable tape recorders suitable for diving operations are commercially available. The input at the electronics package could be sampled at various depths for later spectrum analysis. This procedure would be used to optimize mask/microphone selection.

Electronics improvements include the loading of the I. F. transformers to broaden the band-pass to 20 KHz vice 10 KHz. Active filter band-pass design was matched to the pass characteristics of

the I. F. strip and is easily changed. Since the response of the ratio detector is 20 KHz the I. F. strip is the limting factor in setting beta.

As previously mentioned, local oscillator frequency stability is critical. The use of precision, temperature-insensitive passive elements where possible would enhance frequency stability.

An external charging feature may be easily incorporated by substituting a dual connector for the single connector currently in place. Present configuration requires extraction of the battery pack for re-charging.

Finally, the wire-wrap procedures used in prototype construction should be replaced with printed circuit technology.

The number of recommendations is almost limitless. Included herein are only those which find significant and immediate application in DUCS II.

## LIST OF REFERENCES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